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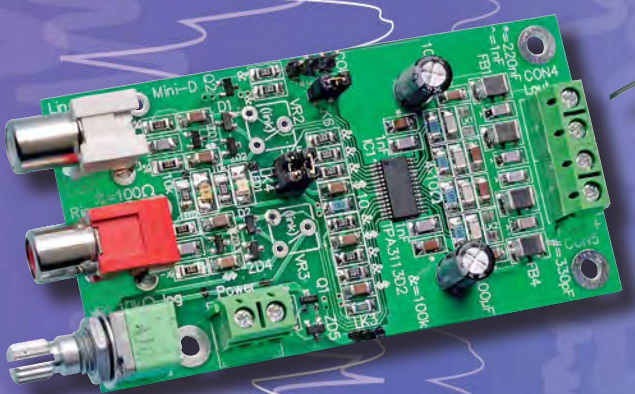

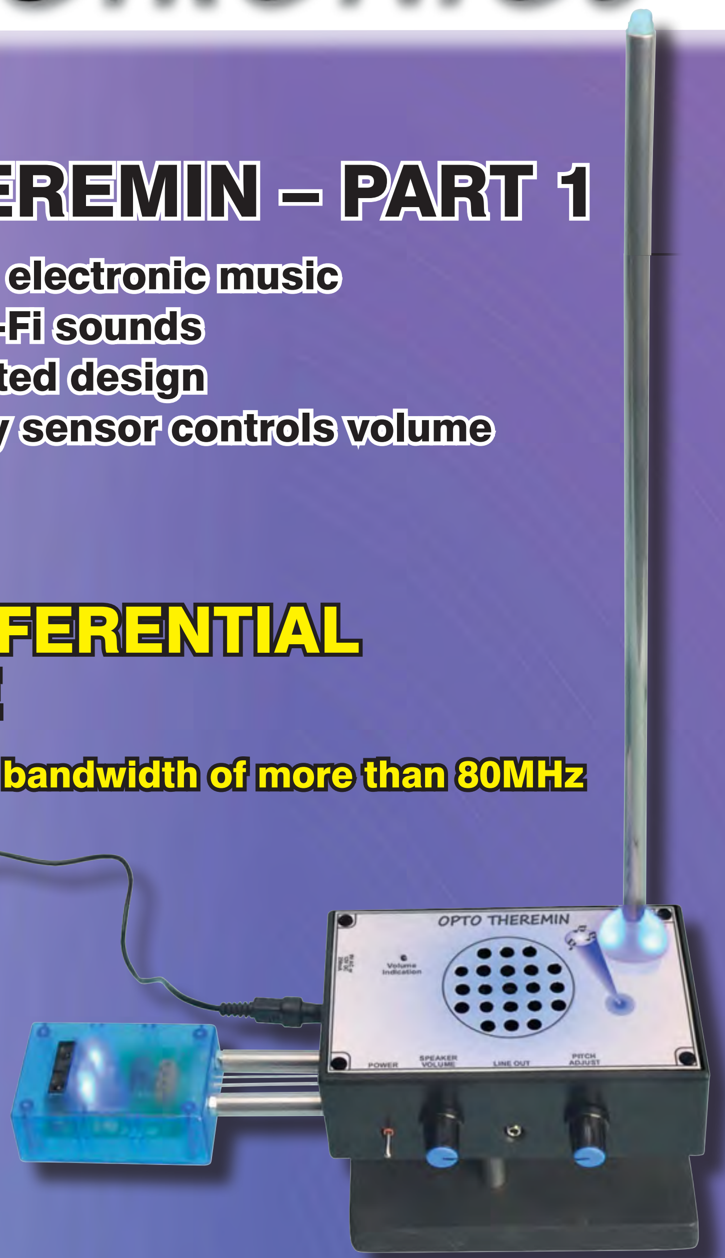

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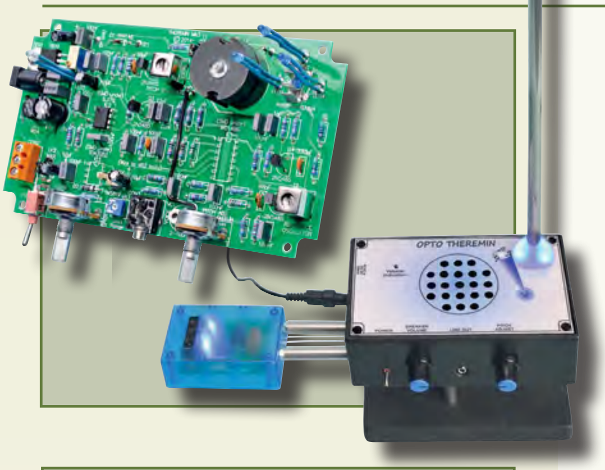

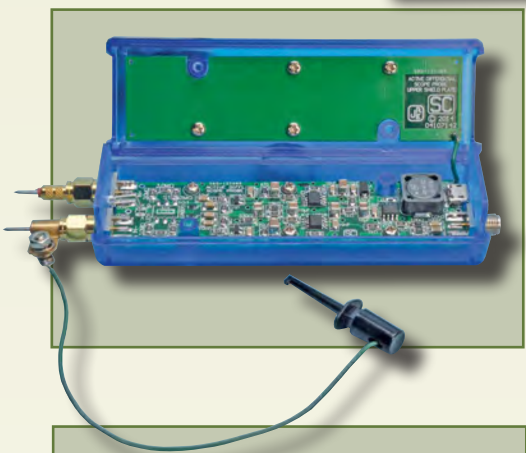

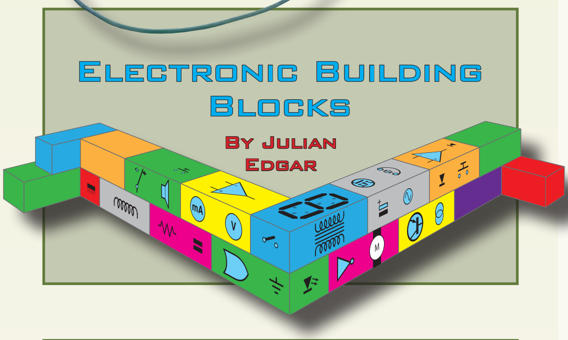

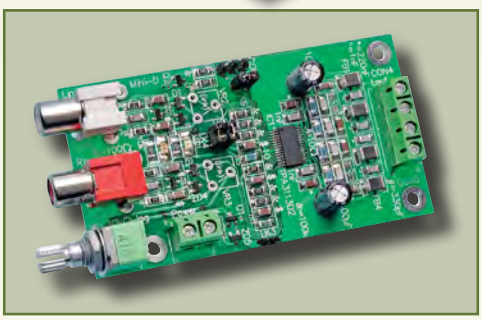

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Our October 2015 issue will be published on Thursday 3 September 2015, see page 80 for details.

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Hot New Products!

Here are a few of the most recent products added to our range. See website or join our email Newsletter for all the latest news.

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EPE EVERYDAY PRACTICAL ELECTRONICS

A little extra for summer

It's summer, so – with luck – you have a bit of extra time to read *EPE* on the beach, in the garden or wherever you relax. With that in mind we have pushed the pagination of this month's issue to 80 pages of electronic fun, learning and inspiration.

Updating a classic

Our first project this month is an update of a classic, if not *the* classic electronic instrument – the Theremin. It's the granddaddy of them all and today remains the only instrument you play without actually touching it! *EPE* has quite a history of publishing Theremin projects. In previous years, Jake Rotham, our *Audio Out* columnist produced a pocket Theremin and a much larger version (you can view some of Jake's work at: **theremin.co.uk**). However, this new design is by regular contributor John Clarke, who has updated the traditional design by using an optical proximity sensor. Even if you don't intend to build one, it's a fascinating instrument to read about.

Trivia

My brief foray into the world of pub-quiz-style questions last month did not exactly result in a tsunami of replies – but we do have a winner! Geoff Theasby spotted that Ian Bell had secreted a Pink Floyd title in *Circuit Surgery*; well almost. 'Careful with that axis' is near enough the name of the early Floyd song 'Careful with that axe, Eugene'. Just to ram home the point, another heading in the same article was 'Any colour you like', also by Pink Floyd. As promised, Geoff's prize is… the honour of winning!

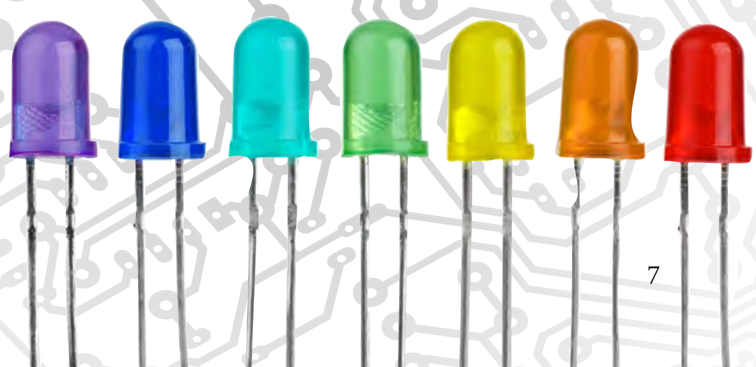
Changes for *PIC n' Mix*

Last, but not least, I have some sad news. Mike Hibbett, our tireless, talented and creative columnist for *PIC n' Mix* has had to give up writing for us. In his own words:

Due to some exciting and positive changes to my fulltime work, I find myself unable to continue writing the PIC n' Mix *column. It's with some sadness that I put down my pen; I've had some great experiences, learned a lot and made many friends. I will continue to participate in the* Chat Zone *forum and support Mike O'Keeffe who takes over the role. See you on the* Chat Zone*!*

It is typical of Mike's modesty that he omitted to point out he has been appointed to a design position in the most prestigious silicon company in the world. The good news is that he will support and help Mike O'Keeffe settle into the role of being our new PIC writer.

We wish Mike (Hibbett) the very best for the future, and thank him for all the wonderful work he has done for *EPE* and the hobbyist community in the UK and further afield.



NEWS

A roundup of the latest Everyday News from the world of electronics

Indoors location – report by Barry Fox

'Indoor location is the last frontier', says Erik Bovee, VP Business Development at Indoors – one of several companies now learning the hard way how hard it is to make satnav systems work without line of sight to satellites in the sky. But whoever cracks the problem will hit a jackpot. Just think how nice it will be to be guided to the correct departure gate at a busy airport, navigate a huge shopping mall or roam a trade exhibition that spreads over many halls, often on several levels.

Useful and legal

The challenge is to provide GPS navigation inside a building, without contravening laws in Europe and the US, which prohibit radio transmission on GPS frequencies. The enabling breakthrough was finalisation of the Bluetooth 4.0 specification five years ago, with the option of 'Smart/Low Energy' signalling. Bluetooth LE sends data in bursts rather than the continuous stream used for conventional Bluetooth. The bursting data can travel further – between 20 and 70 metres, depending on obstructions like walls – and the battery in a beacon lasts at least a year.

Smartphone app and beacons

Broadly, a smartphone app detects Bluetooth from beacons fixed inside a building. The beacons transmit coded fixed-power fingerprint or identification signals and the smartphone detects signal strength (not angle or time of arrival). Comparing the relative strengths of several beacons in the immediate vicinity does the positioning. The position data works with a stored map to display progress along a route, with text or spoken directions.

The beacons can be simple, costing a few dollars each in bulk, and give accuracy of around two or three metres, with a delay of between one and three seconds before the phone delivers a location reading. If the data bursts are more widely spaced; eg, every 500 milliseconds instead of every 100 milliseconds, then the battery can last at least two years. But wider burst spacing increases latency, ie, makes location slower.

With support from Cisco, Indoors is now offering an SDK (software developers kit) which software designers can use to provide apps for mapped locations. A test system at San Francisco Airport helps the visually impaired get around by converting position data into a headphone description of the surroundings or changing the volume of an audible prompt depending on location.

The human problem

It may sound simple but it's not, says Erik Bovee. 'What works for test robots does not work for humans, who don't walk in straight lines and may go backwards or hold their phone upside down.

'The same system could be used to help visitors navigate large trade exhibitions like CES and IFA. We have been talking to Reed Exhibitions. The problem is that exhibitions are only there for a few days, the maps have to be prepared and then it's all over.'

Indoors uses waterproof beacons the size of a matchbox, made by Polish company Kontakt.io. So does another European company Pointr, which was founded in November 2013 after 'trying to do indoor navigation for a client who needed it' and discovering 'there were no sensible solutions available.'

The original beacon standard was set by Apple, but there are now variations on the basic format. Beacons can, for instance, be used for retail advertising, pulling data about a shopper from a cloud store and then pushing individually tailored sales pitch messages such as 'Happy Birthday – we have some special offers for you', based on spend history at the store.

Applications

Another application is 'virtual queuing'. Instead of everyone waiting in line for counter service, beacons push a 'you are next, go to the counter now' message to the appropriate individual – much like the paper ticket systems used in hospital outpatient wards.

Beacons could help emergency services find someone in a crowded building who has called 999.

Indoors demonstrates a fairly crude prototype app in their Vienna labs; the phone displays a simple floor plan map of rooms inside a building, with a blue line tracing the route to be taken and room zones flashing red when entered.

The trials at San Francisco airport have already revealed a practical problem; overloading a blind user with too much acoustic information effectively 'blinds' them to audio help.

BBC unveils finished BBC micro:bit

The BBC has revealed the final design for the micro:bit, a groundbreaking, 29-partner collaboration, including ARM, BBC, element14, Microsoft, Nordic Semiconductor and Samsung. It will provide a pocket-sized computer to every year 7 child in the UK for free. The aim is to to inspire digital creativity and develop a new generation of tech pioneers.

The micro:bit is a pocket-sized, codeable computer that allows children to get creative with technology.

In the 1980s, the BBC Micro introduced many children to computing for the first time. Part of the BBC's 2015 Make it Digital initiative, the BBC micro:bit builds on the legacy of the Micro for the digital age, and aims to inspire young people to get creative with digital; develop core skills in science, technology and engineering; and unleash a new generation of digital makers and inventors.

It measures 4cm by 5cm, will be available in a range of colours, and is designed to be fun and easy to use. Something simple can be coded in seconds – like lighting up its LEDs or displaying a pattern – with no prior knowledge of computing. All that's needed is imagination and creativity.

The BBC micro:bit also connects to other devices, sensors, kits and objects, and is a great companion to Arduino, Galileo, Kano, littleBits and Raspberry Pi, acting as a spring-board to more complex learning.

Final design of the BBC micro:bit, free for all year 7 UK children

Key features

25 red LEDs – lights up, flashes messages and helps users create games.

Two programmable buttons – activated when pressed. Use the micro:bit as a games controller. Pause or skip songs on a playlist.

Motion detector – 'accelerometer' that can detect movement and tell other devices you're on the go. Featured actions include shake, tilt and freefall. Turn the micro:bit into a spirit level. Light it up when something is moved. Use it for motion-activated games.

Built-in compass / ***'magnetometer'*** – to sense which direction you're facing, your movement in degrees, and where you are. Includes in-built magnet to sense certain types of metal.

Bluetooth Smart Technology – connect to the Internet and interact with the world. Connect the micro:bit to other micro:bits, devices, kits, phones, tablets, cameras and everyday objects all around. Share creations or join forces to create multi-micro:bit masterpieces. Take a selfie. Pause a DVD or control your playlist.

Five I/O rings – connect the micro:bit to devices or sensors using crocodile clips or 4mm banana plugs.

Want to help at Bletchley?

Bletchley Park has reported an explosion in visitor numbers. A nice problem to have, but they need more volunteers, in particular stewards. The role involves helping visitors to navigate the once-top-secret site and ensuring it runs smoothly.

So far this year, more than 100,000 people have already visited Bletchley Park; an increase of more than 80 per cent on the previous year. An army of volunteer staff help each and every one of those visitors to get the most out of their day, helping to tell the remarkable story that lives in the very walls of this unique heritage site.

The surge in visitor numbers is due in part to a very special visit a year ago, by the granddaughter and great-niece of twins Valerie and Mary Glassborow, who worked together in Hut 16 towards the end of World War Two. The Duchess of Cambridge visited in June 2014 to formally unveil the rejuvenated Bletchley Park, following the £8 million, Heritage Lottery Fund supported phase one restoration.

Another huge spike in interest followed in November, when the Academy Award-winning film, The Imitation Game, based on the life and work of Alan Turing and starring Benedict Cumberbatch and Keira Knightley, opened in cinemas.

Volunteers at Bletchley Park benefit from working as part of a great team and have access to extensive resources and libraries. Full induction and training is provided. Other roles are also available in Education and Estates. To find out more, please email **volunteer@bletchleypark.org.uk** or call 01908 640404.

85-year hunt may boost next-gen electronics

An international team led by Princeton University scientists has discovered an elusive massless particle theorised 85 years ago. The particle could give rise to faster and more efficient electronics because of its unusual ability to behave as matter and antimatter inside a crystal, according to the research.

The researchers report in the journal *Science* the first observation of Weyl fermions, which, if applied to next-generation electronics, could allow for a nearly free and efficient flow of electricity in electronics, and thus greater power, especially for computers.

Proposed by the mathematician and physicist Hermann Weyl in 1929, Weyl fermions have been long sought by scientists because they have been regarded as possible building blocks of other subatomic particles, and are even more basic than the ubiquitous, negative-charge carrying electron (when electrons are moving inside a crystal). Their basic nature means that Weyl fermions could provide a much more stable and efficient transport of particles than electrons, which are the principle particle behind modern electronics.

IBM reaches 7nm

An alliance led by IBM Research has announced that it has produced the semiconductor industry's first 7nm node test chips with functioning transistors. The breakthrough could result in the ability to place more than 20 billion tiny switches (transistors) on a single fingernail-sized chip.

To achieve the higher performance, lower power and scaling benefits promised by 7nm technology, researchers had to bypass conventional semiconductor manufacturing approaches. Among the novel processes and techniques pioneered by the IBM Research alliance were a number of industry-first innovations, most notably silicon-germanium (SiGe) channel transistors and Extreme Ultraviolet (EUV) lithography integration at multiple levels.

LED there be light

Mark Nelson

Forgive the awful pun (it's probably not even original) but it's a fact that researchers are discovering ways to make even more efficient lighting devices. Meanwhile, a retailer again demonstrates how unbelievably dim it's possible to be. Mark Nelson explains.

ONE OF MY ALL-TIME FAVOURITE quotations comes from the Danish physicist and atom scientist Niels Bohr, who conceded rather regretfully: 'Prediction is very difficult, especially if it's about the future.' Despite this diffidence, he made fundamental contributions to understanding atomic structure and quantum theory, for which he received the Nobel Prize in Physics in 1922. Although he died 53 years ago, his name lives on in the title of the Niels Bohr Institute in Copenhagen, where researchers have announced that LEDs made of nanowires produce more light and consume less energy. Doubtless this is something that Niels himself had not foreseen.

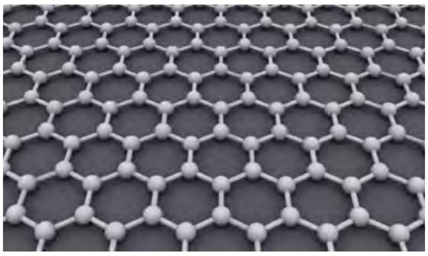
Why nanowires and what are they anyway?

Robert Feidenhans'l, professor and head of the Niels Bohr Institute explains: 'Nanowires are very small – about 10 to 500nm (nanometres) in diameter (one nanometre is a thousandth of a micrometre). When used in LEDs they are made up of an inner core of gallium nitride (GaN) and a layer of indium-gallium-nitride (InGaN) on the outside, both of which are semiconducting materials. The light [produced] in such a diode is dependent on the mechanical strain that exists between the two materials and the strain is very dependent on how the two layers are in contact with each other.'

A company in Sweden produces the nanowires under study, and the Institute's work performed using nanoscale X-ray microscopy indicates that any structural defects in the nano-LEDs significantly affect performance. The new technique enables fine adjustment of the layer structure in the nanowires to eliminate any flaws. The researchers believe these nanowire LEDs may well be ready to use within five years. Professor Feidenhans'l sums up: 'There is great potential in such nanowires. They will provide a more natural light in LEDs and they will use much less power. In addition, they could be used in smart phones, televisions and many forms of lighting.'

World's thinnest light bulb

Are nanowires the way to go? Not necessarily, according to an international research project that has now produced visible light using a filament made of graphene, an atomically thin and perfectly crystalline form of carbon. They attached small strips of graphene to metal electrodes, suspended the strips above the substrate, and passed a current through the filaments to cause them to heat up. The team is led by Young Duck Kim, a potstdoctoral research scientist at Columbia Engineering (New York), assisted by a team of scientists from Columbia, Seoul National University (SNU), and the Korea Research Institute of Standards and Science (KRISS).



Graphene, a one-atom-thick pure carbon material with extraordinary electrical and thermal properties (Image: AlexanderAIUS)

'We've created what is essentially the world's thinnest light bulb,' says Columbia's professor James Hone. 'This new type of 'broadband' light emitter can be integrated into chips and will pave the way towards the realisation of atomically-thin, flexible, and transparent displays, as well as graphene-based, on-chip optical communications.' Co-researcher professor Yun Daniel Park, notes amusingly that they are working with the same material that Thomas Edison used when he invented the incandescent lamp bulb. 'Edison originally used carbon as a filament for his light bulb, and here we are going back to the same element, but using it in its pure form – graphene – and at its ultimate size limit – one atom thick,' he remarks.

Blockheads

No, not the band once fronted by Ian Dury but the company originally known as 'Block and Quayle', now trading under the snappier title of B&Q. If you're planning to make some electrical or other DIY purchases at one of its superstores, I suggest you keep your mobile phone in your pocket at all times. Otherwise you might be banned from shopping there for life. Sounds crazy? – maybe but that's precisely what has happened to two blameless customers.

During May, Raymond Meerabeau visited the B&Q warehouse in New Malden, where he noticed that the cover was missing from a fuse box, exposing mains-voltage wiring.

Being safety-conscious, he used his phone to take photos and politely showed them to a manager. The latter told him aggressively: 'You should not be filming in here and I am going to ban you.' After complaining to the firm's head office, Mr Meerabeau received a reply from B&Q's head of customer care, Gemma Miall-Smithson, banning him from every B&Q in the country and informing him: 'For the avoidance of doubt, please be advised that the ban is a lifetime ban and relates to all our B&Q stores. We trust this clarifies our position.' Which it does and also makes perfectly clear that B&Q haven't a clue about electrical safety, customer relations or how to avoid bad publicity, as this story has now appeared in numerous newspapers and is all over the Internet. A properly trained manager would have thanked Mr Meerabeau profusely and handed him a £25 gift voucher for his public-spiritedness. Now this story will haunt the company, just as Gerald Ratner's 'total crap' description of his firm's products hastened its decline.

Serial offender

James Wright from the town of Thame (small place, but rightly famed for its pet shop) is another customer banned from all B&Q stores for phone 'offences'. He was looking at his phone in there to examine a picture he had taken of a DIY project at home. He was checking out the suitability of some goods in the retailer's Oxford store, but when he returned the phone to his pocket, he was accused of shoplifting after a store detective saw him put something in his pocket. He was apprehended by two 'burly blokes' and subsequently received a banning letter stating: 'You may not, unless notified in writing, visit or enter upon any B&Q stores and/or the above named store. If you do so, you will be regarded as a trespasser entering the premises without lawful authority and legal action may be taken against you.' Utterly charming – and so crass!

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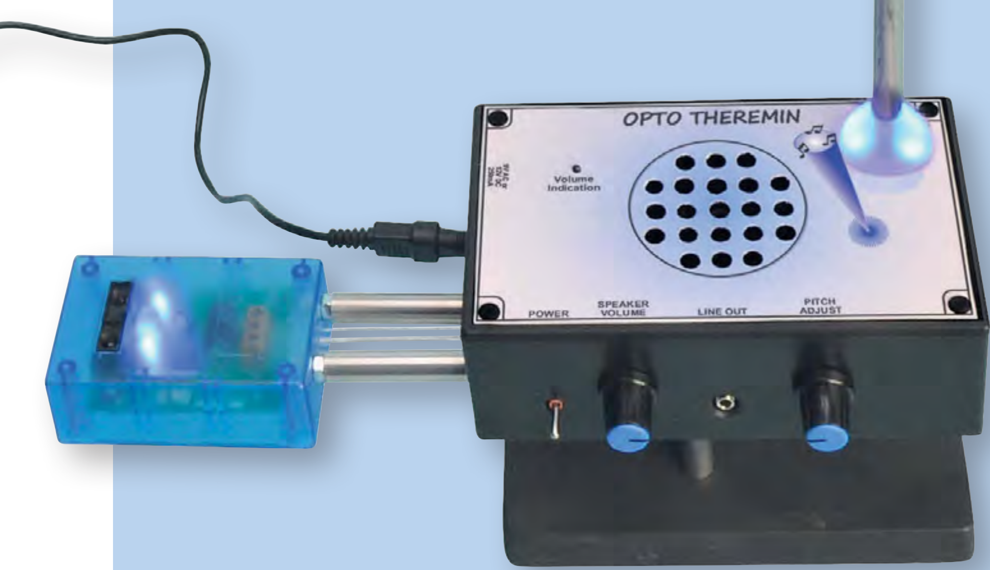
Constructional Project

Create eerie musical sounds with the:

Opto-Theremin – Part 1

Create your own electronically synthesised music or produce eerie science fiction sounds with our new *Opto-Theremin*. This completely new design uses an optical proximity sensor to provide a more effective volume control plate, which adds the possibility of rapid tremolo, while vibrato can be applied in the normal way with the vertical pitch antenna.

By JOHN CLARKE



Unlike conventional Theremins, the new *Opto-Theremin* uses an optical distance sensor to control the volume, making the unit easier to build and adjust. A metal antenna rod is used for pitch control.

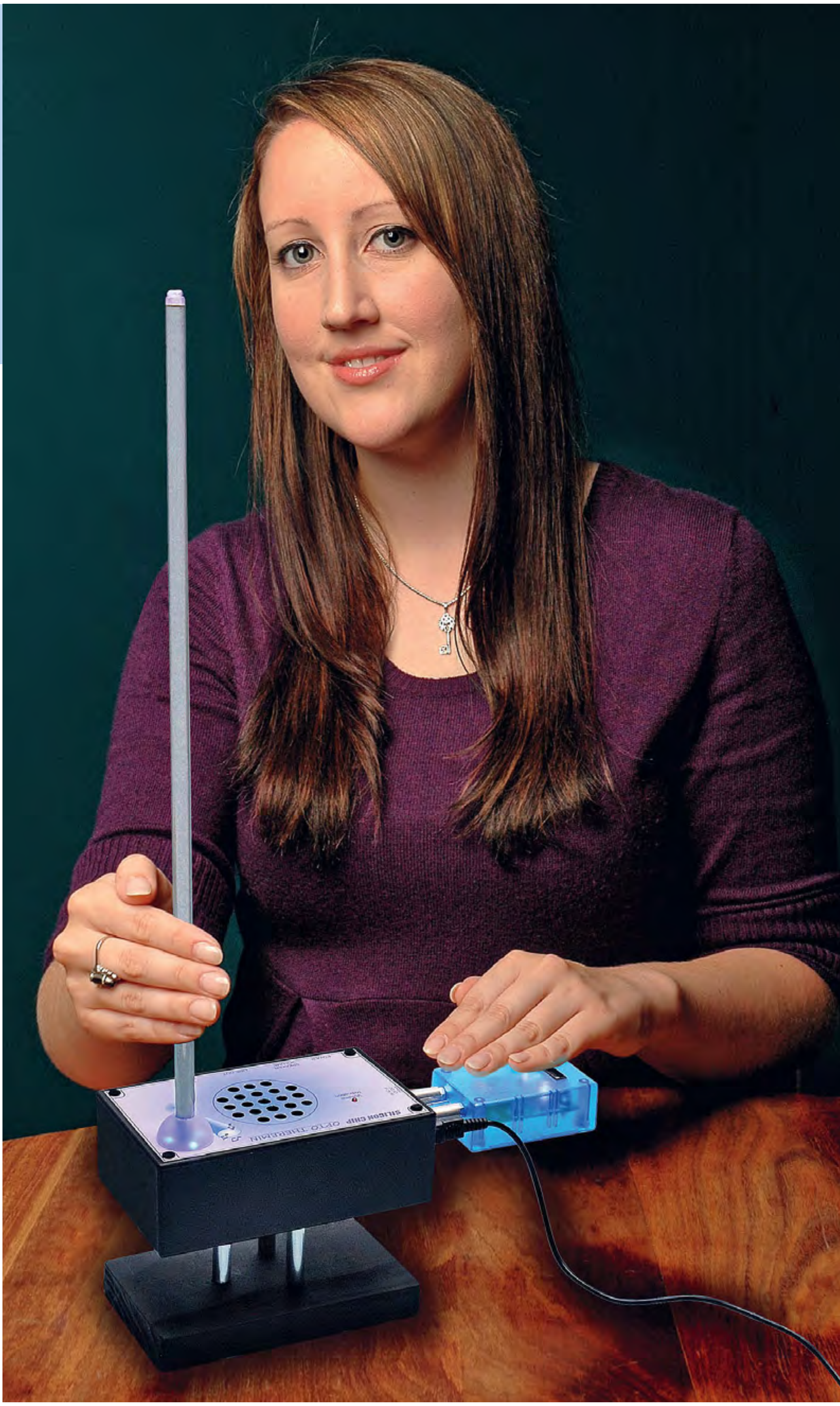
THIS LATEST THEREMIN merges the traditional with the modern. As well as the optical proximity control plate, it includes a touch of 'bling' in the form of blue LEDs and polished aluminium tubes. Even the top of the pitch antenna is illuminated with blue light.

For those who don't know what a 'Theremin' is, it is an electronic musical instrument designed by Leon Theremin in the early 1900s. Pitch and volume are varied by moving your hands near two antennas, and a wide range of tones covering several octaves can be produced. Just do a Google search for 'Theremin' to see a selection of YouTube videos of Theremin performances – but note, they all involve Theremins of largely traditional format.

The Theremin owes its popularity to its extreme versatility and to its unique sound compared to conventional instruments. Even a simple combination of hand movements can lead to interesting effects. Theremin passages can comprise a smooth gliding tone (*glissandi*) or can be separate notes (*staccato*), or a combination of both. It really is a versatile instrument, limited only by the skill of the player.

Our *Opto-Theremin* operates in a radically different manner to traditional Theremin designs. The 'Opto' prefix refers to its use of an optical volume control and to the blue LEDs that add visual interest – the 'bling'.

Before anyone starts worrying that our new *Opto-Theremin* may have lost its heritage, be assured that it sounds just like a traditional Theremin and is played in exactly the same way. For example, the *Opto-Theremin* still has a vertical antenna for pitch control just like a traditional Theremin, whereby the right hand is moved horizontally to change pitch.



The big difference compared to a conventional Theremin is the volume control. As with the original, the left hand is moved vertically to control the volume, but this movement is sensed using an optical proximity sensor rather than the traditional horizontal loop-shaped antenna.

Why use optical sensing?

This solves a number of problems. Traditional Theremins use RF (radio frequency) oscillators to feed the antennas for both pitch and volume control. Without careful tuning, there can be all sorts of interactions between the volume and pitch oscillators, leading to unwanted 'squawks' in the sound, or pitch changes when the player is only trying to adjust the volume.

By using optical sensing for the volume control instead, there's no chance of any interaction with the pitch control circuitry. Additionally, the volume action is always predictable and does not drift with temperature changes. Plus, it makes the set-up procedure much easier.

We're still mixing two high-frequency oscillators to produce the audio signal, as this results in sounds with the required waveform to imitate musical instruments, such as a cello. So although this new *Opto-Theremin* has a different method for volume control, it still uses RF techniques to generate the pitch, allowing it to produce the classic Theremin sound.

Features

In order to play a Theremin, the musician must be able to accurately position one hand near the antenna, to produce the required pitch. The generated tone has to be set 'by ear', just as for a violin or a trombone. This is because the Theremin does not have a fixed set of notes, but instead delivers a continuous range of tones over several octaves.

Correct linearity of pitch variation in response to hand movement is a critical feature of the design. In this case, 'linearity' means that there is a similar range of hand movement for each octave. It's important that no

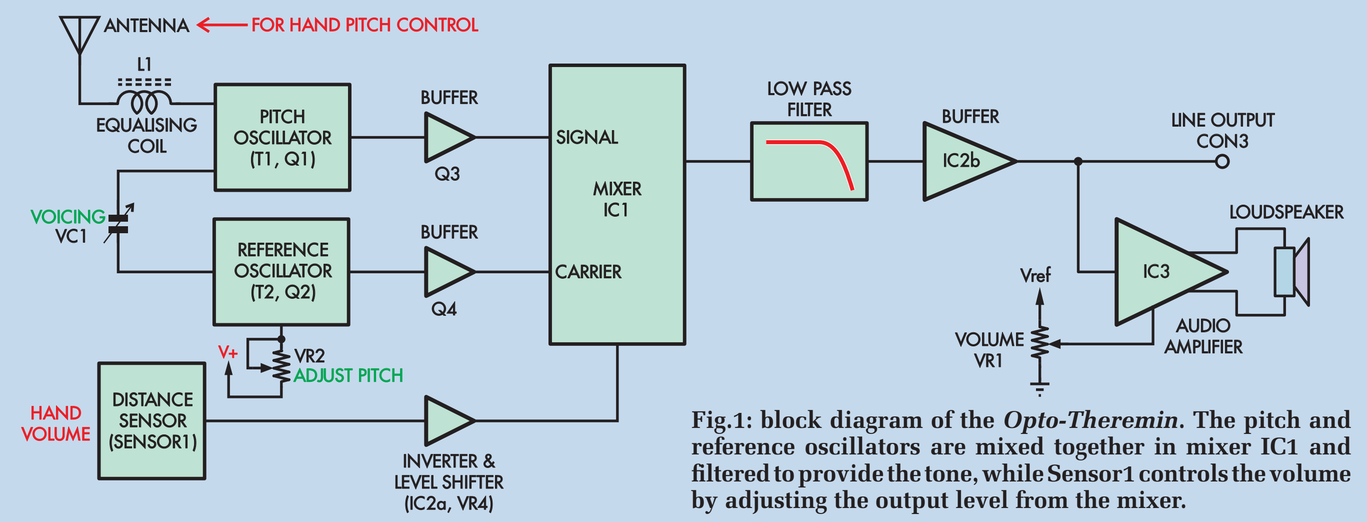


Fig.1: block diagram of the *Opto-Theremin*. The pitch and reference oscillators are mixed together in mixer IC1 and filtered to provide the tone, while Sensor1 controls the volume by adjusting the output level from the mixer.

octave is compressed into a very small hand movement range, as this would make the instrument difficult to play.

The *Opto-Theremin* is designed to avoid this and it includes a test circuit to assist in correctly adjusting the linearity. An adjustment is also included to modify the tonal quality or 'voice' of the *Opto-Theremin*. This allows it to be adjusted from producing a sinusuoidal (or pure) tone through to a sound that's reminiscent of a cello at low frequencies and a soprano voice at higher frequencies. In addition, an externally adjustable pitch control provides compensation for changes in pitch due to the unit's location and its surroundings and/or due to temperature variations.

The unit contains an in-built amplifier and loudspeaker, but it also has a 'Line Out' socket on the front panel so it can be connected to an external amplifier and loudspeaker system. The loudspeaker volume is independently adjustable so it can be silenced when using an external amplifier, or alternatively, used as a monitor speaker during on-stage performances.

Main features

- External pitch adjustment control
- Linear pitch change with hand movement over four octaves
- Linear volume control with hand movement
- Adjustable hand volume range
- Voicing adjustment (internal)
- Integral loudspeaker with volume control
- Minimal pitch drift during warm-up
- No volume control drift during warm-up
- 9VAC or 12V DC operation @ 250mA (eg, from AC plugpack or 12V battery)
- Line output level: 250mV RMS
- Frequency range: <40Hz to >5kHz

Appearance and controls

As shown in the photos, the *Opto-Theremin* is housed in two plastic cases, one to accommodate the main PCB (and support the pitch antenna) and a smaller one to house the distance sensor PCB for the volume control. They are connected by threaded rods housed within aluminium tubes and the whole assembly mounts on a timber pedestal via another set of aluminium tubes and rods.

The vertical pitch antenna is also made from aluminium tube, and is easily detached for transportation. The volume control box is translucent and lights up during operation to look 'cool'. A translucent dome at the base of the pitch antenna is also lit using blue LEDs, while a separate blue LED illuminates the transparent cap at the top. These blue LEDs not only give the *Opto-Theremin* an impressive appearance, but also reflect from the player's hands when the instrument is being played, for even greater visual effect.

The three external controls (power, volume and pitch) are arranged along one side of the case, together with the line output socket. Power can come from a 9VAC supply or from a mains-derived 12VDC supply or battery. **Note that a switch-mode DC supply (eg, a switch-mode DC plugpack) is not suitable for use with the *Opto-Theremin*.** That's because noise from a switch-mode supply would find its way into the two onboard oscillators and upset the operation.

Operating principles

Fig.1 shows the block diagram of the *Opto-Theremin*. It uses two oscillators: (1) a pitch oscillator and (2) a reference oscillator. Both oscillators are set to run at close to 455kHz. The reference oscillator includes pitch adjustment VR2, to precisely trim the frequency.

While the reference oscillator basically runs at a fixed frequency, the pitch oscillator is varied via the attached antenna. Any hand movement adjacent to the pitch antenna alters its coupling to ground and this changes the frequency of oscillation.

Both oscillator outputs are buffered to isolate them from the following mixer stage, an MC1496 balanced modulator (IC1). As shown, the signals are fed to the SIGNAL and CARRIER inputs of IC1.

Its output comprises several frequencies, including the sum and difference frequencies of the reference and pitch oscillators.

If the two oscillators are almost at the same frequency, eg, 455kHz and 454kHz, then the sum of the two frequencies will be 909kHz while the difference frequency will be 1kHz. The low-pass filter on the mixer's output removes all frequencies above 3.3kHz, leaving only the difference frequency; in this case, 1kHz.

The resulting 1kHz audible tone is then fed to unity-gain op amp stage IC2b, which buffers it and provides the line output signal. This also drives a small internal power amplifier (IC3) and loudspeaker.

So far, we haven't mentioned the equalising coil that's connected between the pitch antenna and the pitch oscillator. This vastly improves the linearity of the pitch oscillator's response as it changes frequency due to hand movements near the antenna. Without it, relatively small hand movements would cause large frequency changes at the higher octaves.

The equalising coil works by forming a tuned circuit in conjunction with the capacitance of the antenna. Its resonant frequency is set to just below the pitch oscillator's frequency by its 9mH inductance and the antenna's ~14pF capacitance. Moving a hand closer to the antenna increases this capacitance, thereby reducing L1's resonant frequency.

In practice, changes to the equalising coil's resonant frequency will be much greater than any corresponding frequency changes in the pitch oscillator. This is because hand capacitance effects of just few picofarads will have a far greater effect on the antenna's 14pF capacitance (and hence the resonant frequency of the equalising coil) than on the much larger 220pF capacitor that's in parallel with the 560µH pitch oscillator coil (both contained within a 455kHz IF transformer).

So, with the equalising coil, hand capacitance changes have a greater effect on the pitch oscillator for hand movements further away from the antenna than closer in. This nonlinearity counteracts the non-linearity of the pitch oscillator's sensitivity to capacitance changes and results in the required linear response.

For further information on this, see: **www.element14.com/community/thread/1802/l/theremin-linearity**

Trimmer capacitor VC1 adjusts the coupling between the pitch and reference oscillators. This is the 'Voicing' adjustment, and it affects the waveshape of both oscillators due to intercoupling, thus also affecting the output waveform shape.

In practice, it's just a matter of setting VC1 to obtain the required sound from the *Opto-Theremin*.

Volume control

As mentioned, we use an optical distance sensor (made by Sharp) for the volume control. It comprises an infrared transmitting LED, a receiving lens and a sensor array. The LED and the receiving lens are spaced about 20mm apart, while the sensor array is a charge-coupled device (CCD) consisting of numerous light sensors arranged in a single row.

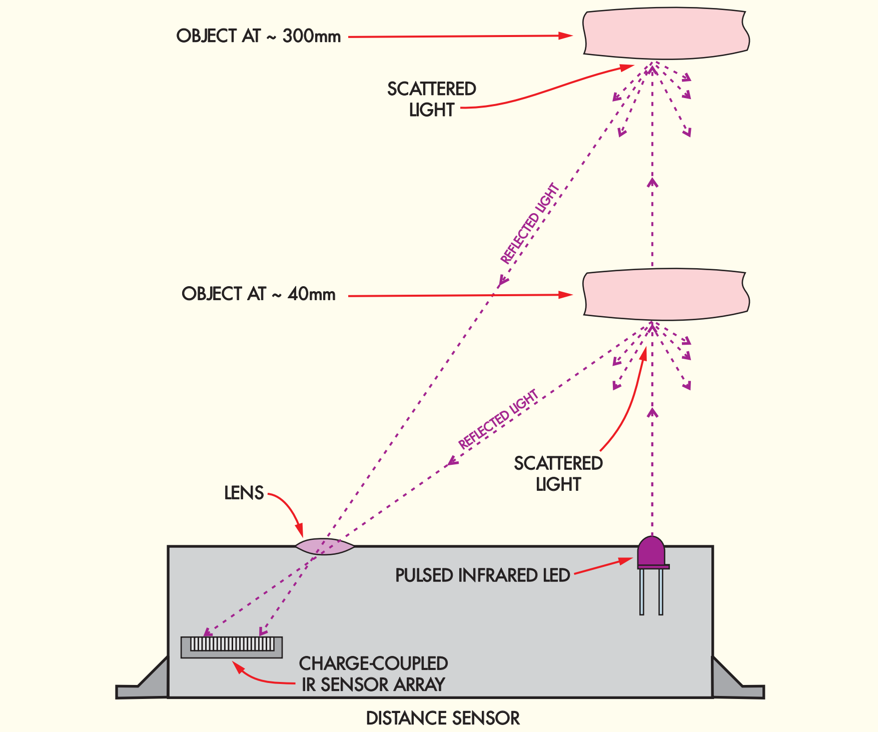


Fig.2: how the optical distance sensor works. As the object moves away from the pulsed infrared LED, the angle of the reflected light passing through the lens changes and this changes the position of the light spot focussed onto a charge-coupled device sensor array (or CCD).

In operation, the LED is pulsed so that it produces high-intensity flashes of infrared light focused as a small dot. If an object is within the sensor's range of measurement, the infrared light will be reflected and some of it focused by the lens.

If the reflecting object has an uneven surface, the infrared light will tend to be scattered – see Fig.2. However, part of the light will be reflected back to the lens, which then focuses it on the CCD.

The exact position of the light spot on the CCD will depend on several things: (1) the spacing between the IR LED and the lens; (2) the distance between the focal point of the lens and the CCD's light-sensitive surface; and (3) the distance from the reflecting object to the sensor.

The first two distances are fixed by the sensor itself, leaving the distance between the sensor and the reflecting object as the variable.

If the object is close to the sensor, the reflected light will be focussed towards the outside edge of the CCD. However, as the object moves further away, the reflected light angle becomes progressively shallower. As a result, the reflected light progressively moves towards the other end of the CCD. The sensor includes circuitry to detect where the light is focussed on the CCD and processes this information to produce a voltage output that varies with distance.

Note that the object does not need to be perfectly flat or parallel to the sensor. The sensor will detect the object as long as there is sufficient scattered light from the object to reach the lens.

Sharp makes several different versions of the distance sensor, each with different optics that set the range of distance measurements. The *Opto-Theremin* uses the GP2Y0A41SK0F sensor, which has a range of 40-300mm. For further information on this device, refer to the data sheet at **www.sharp.co.jp/products/device/doc/opto/gp2y0a41sk_e.pdf**

The output from the distance sensor drives IC2a which inverts and level shifts the signal. IC2a's output then supplies bias current to mixer stage IC1, to control the volume. Inverter stage IC2a is necessary because the output voltage from the sensor reduces with distance, but we want the volume

Background to the Theremin

In 1919, Russian Physicist Lev Termen (or Leon Theremin as he is called in the West) invented an electronic musical instrument called the 'Theremin'.

At that time, the Theremin was innovative and unique in the musical world and was essentially the first electronic instrument of its kind. Playing it relied solely on hand movements in the vicinity of two antennas to control two electronic oscillators – one antenna to vary the pitch of the sound and the other to change the volume.

The Theremin was subsequently further developed and manufactured by the Radio Corporation of America (RCA) around 1929. General Electric (GE) and Westinghouse also made Theremins in the 1920s. However, the number of units produced was quite modest, totalling about 500.

Today, the Theremin is hailed as the forerunner to modern synthesised music and was instrumental in the development of the famous Moog synthesisers. There is also a website devoted to Theremins (**www.thereminworld.com**). Because of its unique sound, it has been popular with music producers for both film and live performances. The sound is ideal for setting the scene for supernatural events and for close encounters with extraterrestrial beings in science fiction movies.

A Theremin was used to produce background music in the feature film 'The Ten Commandments' by Cecil B DeMille (1956). Its eerie sounds have also made it ideal for science fiction movies such as 'The Day the Earth Stood Still' (1951), 'Forbidden Planet' (1956) and 'Mars Attacks!' (1996).

The Beach Boys also used an instrument similar to the Theremin – called an Electro-Theremin (also named a Tannerin) – in their 1966 hit, 'Good Vibrations'. More information on Theremins is available at **www.thereminworld.com/**

Theremin models

EPE has produced previous designs for home construction: the *Mini Theremin* in May/June 2008 and the *Mk2 Theremin* in March 2011. Last, but not least, *EPE's* Audio Out columnist offers coventional Theremin kits at: **www.theremin.co.uk**

to increase as the hand is moved further away (ie, upwards).

Circuit details

Fig.3 shows the full circuit of the *Opto-Theremin*. As well as the distance sensor (SENSOR1), it uses three low-cost ICs (IC1-IC3), four JFETs (Q1-Q4), several coils and sundry other parts.

Both the pitch and reference oscillators utilise pre-wound 455kHz IF (intermediate frequency) transformers (T1 and T2), as commonly used in AM radio tuners. Each of these stages is connected as a common-drain Hartley oscillator, with T1 and Q1 forming the pitch oscillator, and T2 and Q2 making up the reference oscillator.

T1 has a tapped primary winding with a parallel-connected capacitor to form a tuned circuit. Its resonant frequency can be varied using a ferrite slug which screws into the core. Q1 drives a portion of the tuned circuit winding via the tapping at pin 2, while the signal at the top of the tuned winding is coupled to the self-biased gate of Q1 via a 68pF capacitor. This arrangement provides positive feedback to maintain oscillation at the tuned frequency.

The second winding inside T1, at pins 4 and 6, provides a low-impedance output signal. This signal is fed to the gate of JFET Q3 via a 330pF capacitor. Q3 is wired as a source-follower stage, buffering the signal from T1 and feeding it to pin 1 (SIG IN+) of mixer IC1. Current is fed to Q1's drain via a 680Ω resistor connected to the 9V DC supply rail, while Q3's drain current is set by a 100Ω resistor to ground.

The reference oscillator is very similar to the pitch oscillator, the difference being that JFET Q2 is powered via 1kΩ potentiometer VR2 and a 220Ω resistor. VR2 varies Q2's drain-source current to provide pitch adjustment since altering this current affects Q2's gate-source capacitance. This in turn alters the reference oscillator's tuned frequency. Q4 buffers the signal from the reference oscillator, feeding it to pin 8 (CARRIER IN+) of IC1.

Equalising coil L1 is connected directly to pin 1 of T1 by placing jumper link LK1 in its NORMAL position. Moving LK1 to the TEST position means that the equalising coil is in series with a 100kΩ resistor.

Diode D1 connects to the junction of the equalising coil and the 100kΩ resistor, while its cathode goes to test point TP1. In test mode, the equalising coil is sufficiently isolated from the pitch oscillator to allow the resonance of the coil and antenna to be monitored by a DMM set to read DC volts, connected between TP1 and TP GND.

In operation, the DMM filters the rectified RF signal from D1 due to both lead capacitance and internal capacitance, and it discharges this stray capacitance via its own loading. Once the DMM is in place (and LK1 set to TEST), the slug in T1 is adjusted to alter the frequency of the pitch oscillator to give the lowest voltage reading. This sets the pitch oscillator to the resonant frequency of the equalising coil and antenna. The frequency is then adjusted slightly away from this resonance point.

Mixer stage

As mentioned, the signals from JFET buffer stages Q3 and Q4 are applied to pins 1 and 8 of mixer IC1 via 1nF capacitors. The signal level applied to pin 1 is around 180mV, while the level applied to the carrier input at pin 8 is reduced to around 50mV by the resistive divider at Q4's source, preventing carrier overload.

The signal inputs at pins 1 and 4 and the carrier inputs at pins 8 and 10 are all DC biased from a voltage divider connected across the 9V supply. This divider comprises the 1.2kΩ, 820Ω and 1kΩ resistors and each input is connected to the divider via a 1kΩ resistor.

Note that the SIG IN– and CARRIER IN– inputs (pins 4 and 10) are only DC biased, with any AC shunted to ground via 100nF capacitors.

The 680Ω resistor between pins 2 and 3 of IC1 sets the gain of the mixer, while the bias voltage applied to pin 5 (from IC2a) sets the signal level at the two output pins (6 and 12). As shown, these outputs are biased using 2.2kΩ pull-up resistors (to the 9V rail) and filtered using 22nF capacitors to remove ultrasonic signals.

Unity-gain op amp stage IC2b buffers the low-pass filtered audio signal from pin 6 of IC1. The signal is AC-coupled via a 100nF capacitor to IC2b's non-inverting input (pin 5), while a resistive divider consisting of two 100kΩ resistors across the 9V supply biases this input to 4.5V.

IC2b's output appears at pin 7 and is fed to the Line Out socket (CON3) via

THE OPTO THEREMIN

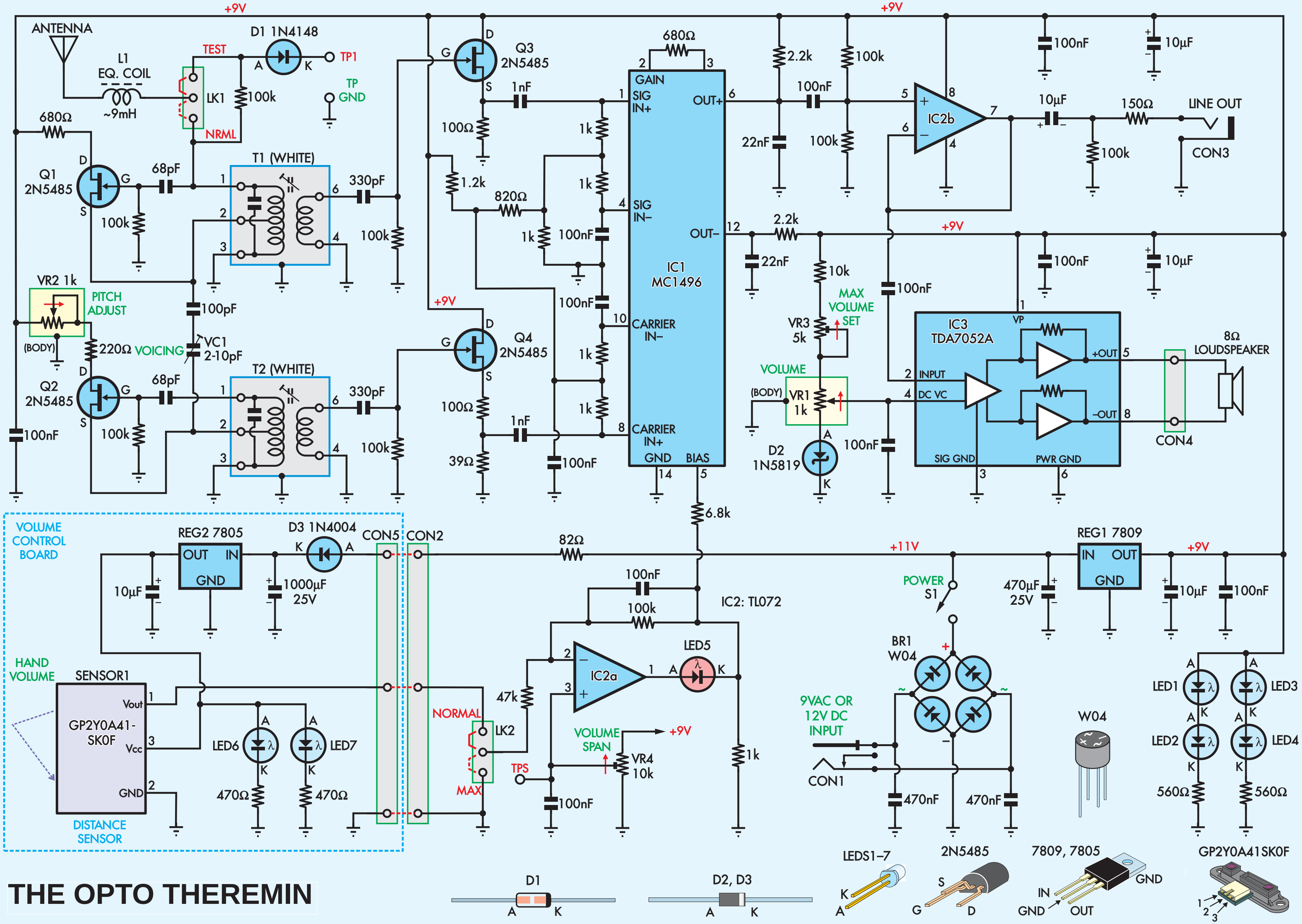
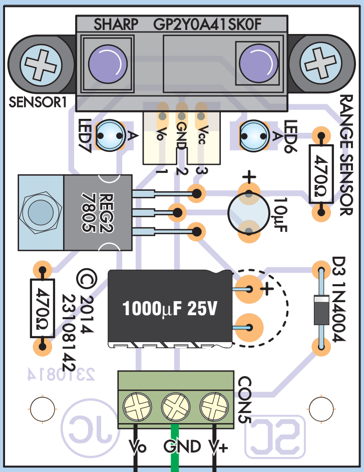


Fig.3: the complete circuit diagram for the *Opto-Theremin*. JFET Q1 and transformer T1 form the pitch oscillator, while Q2 and T2 form the reference oscillator. Their outputs are buffered by Q3 and Q4, and mixed in IC1. Pin 6 of IC1 then drives the line output socket via buffer stage IC2b, while IC2b drives audio amplifier stage IC3. Sensor 1 is the optical distance sensor. Its output is buffered and inverted by IC2a, which then drives the BIAS input of IC1 to control the volume.



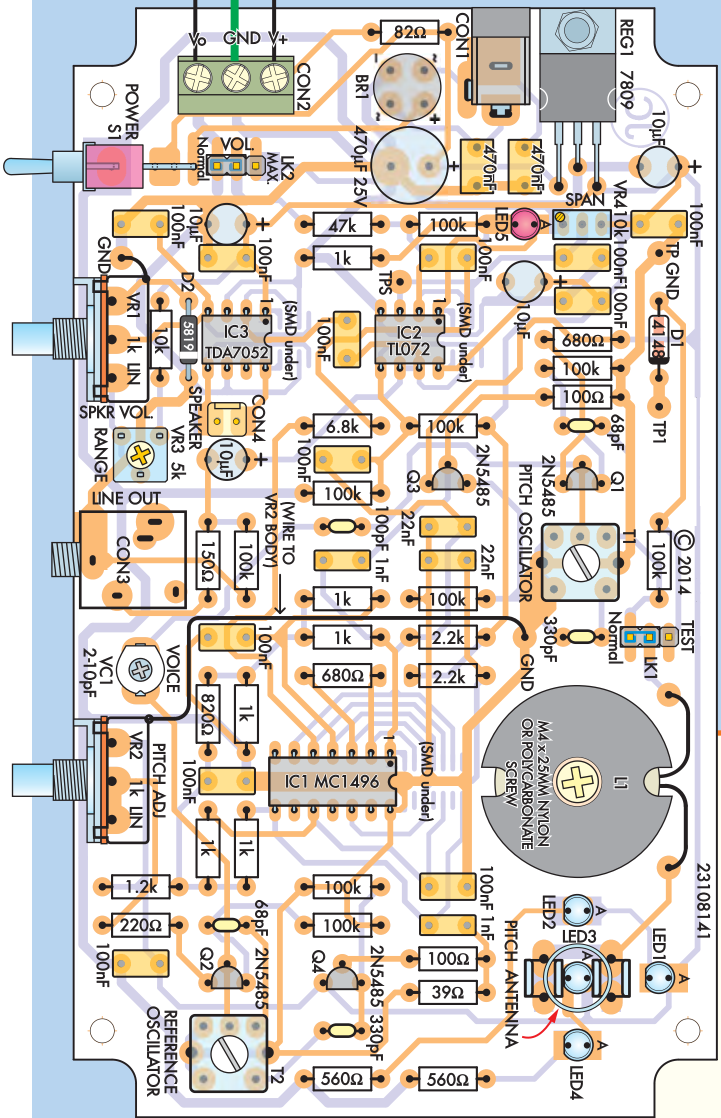
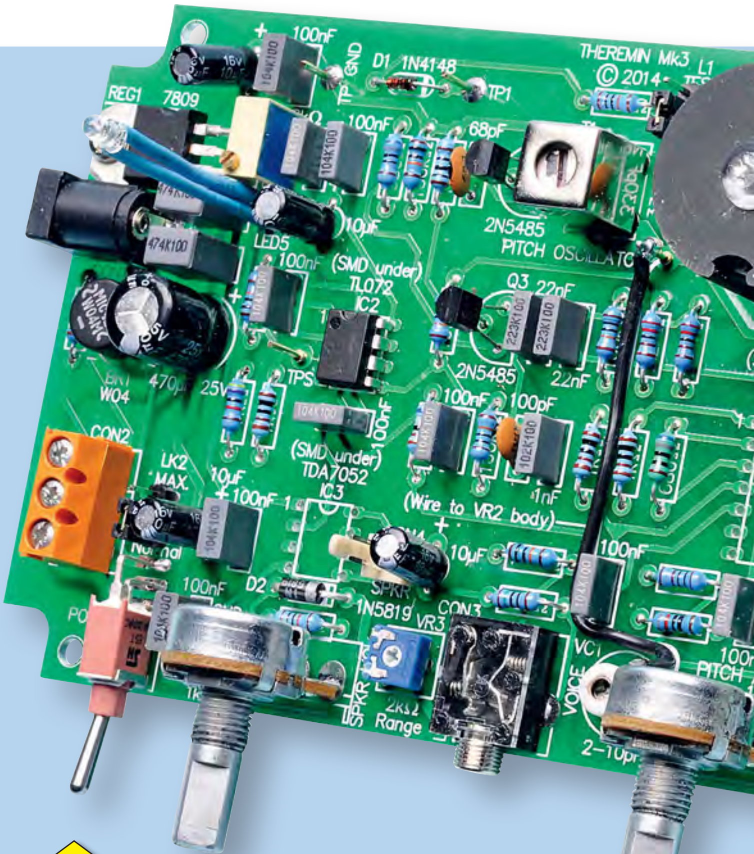
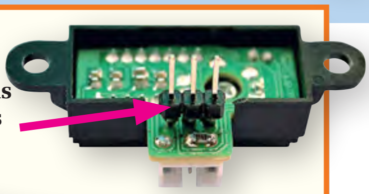


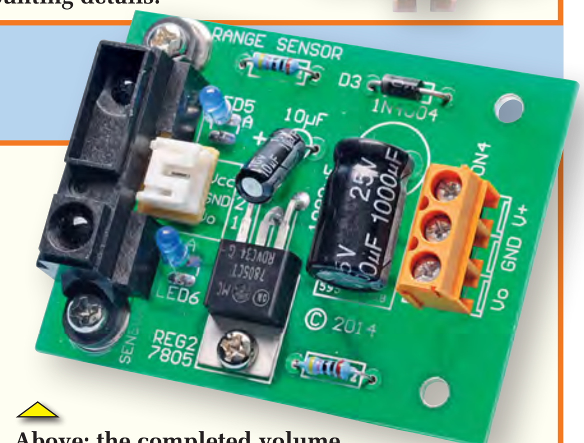
Fig.4: install the parts on the two PCBs as shown in this parts layout diagram, starting with the main PCB assembly as shown below. If you are using SMD ICs for any of IC1-IC3, then these should be installed on the back of the main PCB, as shown in Fig.5. Note that equalising coil L1 must be secured to the PCB using an M4 × 25mm nylon or polycarbonate screw and nut (*do not* use a metal screw!).



The main PCB has been designed to accept either DIP or SOIC (surface-mount) ICs (IC1 and IC3 are SOICs on this assembly, while IC2 is a DIP IC).

Right: a 3-pin header is soldered to the distance sensor's output terminals before installing it on its PCB – see Fig.6 for the mounting details.



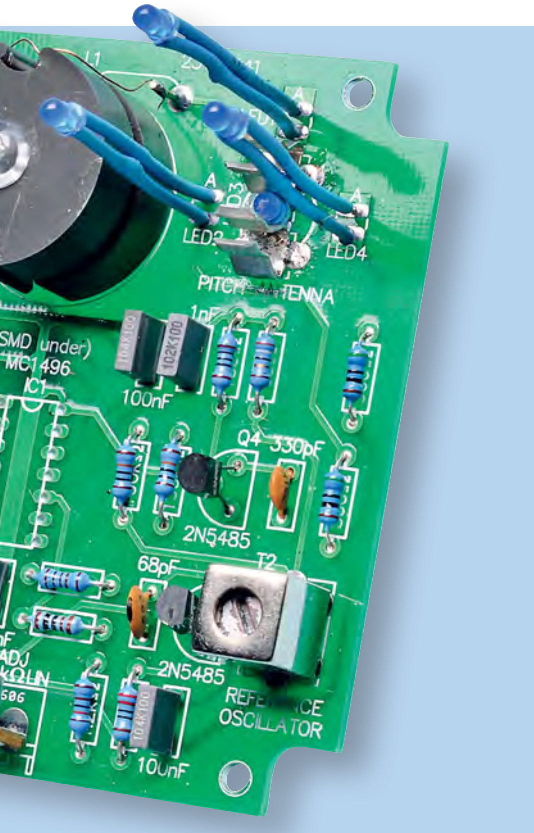


Above: the completed volume control PCB. Note how the two electrolytic capacitors are bent over so that they later clear the case lid.

a 10µF coupling capacitor (to remove the 4.5V DC bias voltage) and a 150Ω resistor. The 150Ω resistor isolates the op amp from any capacitive loads, preventing oscillation.

IC2b's output also feeds power amplifier IC3, a 1W bridge-tied load (BTL) amplifier. Its volume is controlled by a DC voltage at pin 4, with a range of about –70dB to +35dB for 0.4-1.2V.

Volume control pot VR1 is connected in series with trimpot VR3 and a 10kΩ resistor from the 9V supply, with VR1 being the volume control and VR3 being the maximum volume preset.



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VR3 allows the top of VR1 to be adjusted from 0.75-1.0V, giving a maximum gain between about –20dB and +20dB. In practice, VR3 is set so that the loudspeaker produces sufficient volume without gross distortion.

The bottom end of VR1 connects to ground via Schottky diode D2. This provides a fixed bias of approximately 0.2V at the bottom of VR1 and is necessary to set the minimum volume level.

Optical volume control

The Sharp GP2Y0A41SK0F distance sensor (SENSOR1) forms the heart of the optical volume control circuit. Its output at pin 1 varies from about 0.4V when the hand is 300mm above the sensor, to about 2.8V at 40mm. The sensor's output is non-linear and must be inverted and level shifted using op amp IC2a to derive the correct volume control function to drive the bias input (pin 5) of mixer stage IC1.

As shown on Fig.3, the sensor's output is fed to the inverting input of IC2a via LK2. IC2a operates with a gain of just over –2, as set by the ratio of the 100kΩ and 47kΩ feedback resistors. IC2a's non-inverting input (pin 3) is biased to about 1.7V by trimpot VR4 and this offsets the output by 1.7V × the non-inverting gain, ie 1.7V × (1+ 100kΩ/47kΩ) = 5.3V. VR4 allows the volume control range to be set to suit the degree of hand movement.

IC2b is configured in a rather unusual way, with its output driving a red (or green) LED (LED5) and a 1kΩ resistor to ground. The arrangement ensures that the output at LED5's cathode can swing all the way down to 0V. This is necessary because IC2a's output can only go down to 1.8V (it's a TL072) and we need 0V to set the minimum bias on pin 5 of IC1.

So why not use an op amp that can swing down to 0V, such as an LMC6482 or LM358? The answer is that these aren't tolerant of RF signals and produced high-frequency noise in this circuit, even with extra compensation and filtering. The TL072 doesn't have this problem. In addition, LED5 acts as a volume indicator, displaying full brightness at maximum volume and dimming down as the volume is reduced.

The output from LED5 drives the bias input of IC1 via a 6.8kΩ resistor. With 0V output, the lack of bias completely shuts down any signal at IC1's output to provide full attenuation.

The maximum output from IC2a is around 7V. So after taking the LED voltage drop into account, the maximum voltage that can be applied to IC1's bias input is about 5.2V, sufficient to give full volume.

Link LK2 is included so that the distance sensor can be bypassed. When it's moved to the MAX position, pin 2 of IC2b inverting amplifier is tied to 0V via a 47kΩ input resistor. As a result, IC2b's output goes high and the distance sensor no longer has any effect, making pitch adjustments easier.

Power supply

As stated, power for the circuit is derived from a 9VAC plugpack or from a 12V DC linear (non-switch-mode) supply. RF is filtered from the incoming AC (or DC) rails by 470nF capacitors, while BR1 full-wave rectifies the AC supply. BR1 also makes the unit insensitive to DC polarity. A 470µF capacitor filters the resulting DC, while regulator REG1 provides the 9V rail to power most of the circuit.

A 5V supply rail for the distance sensor is derived via diode D3 and regulator REG2. D3 provides reverse-polarity protection, while the following 1000µF filter capacitor is necessary to supply the peak current for the pulsed infrared LED inside the sensor. An 82Ω resistor in series with the 11V supply input limits the peak charging current into the 1000µF capacitor. This prevents unwanted noise in the output due to the pulsing of the IR LED in the sensor.

LEDs 5 and 6 illuminate the area adjacent to the volume sensor with blue light when power is applied. A 470Ω resistor in series with each LED provides current limiting.

Construction

Virtually all the parts for the *Opto-Theremin* are mounted on the two PCBs. The main PCB, available from the *EPE PCB Service*, (coded 23108141) is double-sided and measures 147 × 85mm, while the volume control PCB, also available from the *EPE PCB Service,* (coded 23108142) is single-sided and measures 61 × 47mm. Fig.4 shows the parts layout for both boards.

Start by assembling the main PCB. This board has been designed to accept either DIP or SOIC (surface-mount) packages for IC1-IC3. DIP package ICs are installed on top of the PCB, while SOIC package ICs go on the underside, as shown on Fig.5.

DIP ICs are somewhat easier to install, but many types are now difficult to obtain in this package, especially the MC1496 and TDA7052. An SOIC package is still quite easy to solder though, even though its pins are closer together.

If using one or more SOIC packaged (SMD) ICs, then these should be installed first (see Fig.5). Begin by placing a tiny amount of solder on one of the corner pads, then coat the remaining pads with some no-clean flux paste. That done, place the IC in position (with the correct orientation) and hold it in place using tweezers.

Now solder the relevant corner pin to its pad, then check that the IC is correctly positioned, with all pins centrally located on their pads. If the IC needs adjustment, reheat the soldered pin and slide the IC to its correct position.

Once it's correct, solder the remaining pins but don't worry about solder bridges between pins during this procedure. Once all the pins have been soldered, you can remove any excess solder using solder wick.

Parts List

Main Theremin Section

1 double-sided PCB with plated-through holes, available from the *EPE PCB Service*, code 23108141, 147 × 85mm
1 UB1 plastic utility box, 158 × 95 × 53mm
1 front panel label, 149 × 87mm
1 9VAC 250mA plugpack
1 PCB-mount DC socket (inner diameter to suit plugpack) (CON1)
1 3-way PCB-mount screw terminal block with 5.08mm pin spacing (CON2)
1 PCB-mount 3.5mm stereo switched socket (CON3)
1 2-way polarised header, 2.54mm spacing (CON4)
1 SPDT miniature PCB-mount toggle switch (S1)
1 75mm 8Ω loudspeaker
2 1kΩ linear 16mm potentiometers (VR1,VR2)
2 knobs to suit potentiometers
1 5kΩ horizontal trimpot (VR3)
1 10kΩ multi-turn top adjust trimpot (VR4)
2 2nd IF coils (white) (T1,T2) (can be bought in a set of IF coils from Jaycar Cat LF-1050. Two sets required)
1 potcore pair and bobbin (L1) (Jaycar LF-1060 cores/LF-1062 bobbin, Altronics L 5300 cores/L 5305 bobbin)
2 M205 PCB-mount fuse-clips for antenna connection
1 2-way polarised header plug, 2.54mm spacing, with crimp pins
2 3-pin headers with 2.5mm spacing (LK1,LK2)
2 jumper shunts (for LK1 and LK2)
1 M4 × 25mm nylon or polycarbonate screw (to secure L1)
1 M4 × 10mm nylon or polycarbonate screw (for top of pitch antenna)
2 4mm ID nylon or polycarbonate washers (spacer for L1)
3 M4 nylon or polycarbonate nuts (to secure L1 and for top of pitch antenna)
3 M3 × 6mm machine screws
2 M3 × 10mm machine screws
3 M3 nuts
2 M3 × 9mm tapped spacers
1 100mm length of medium duty hookup wire (to earth VR2)
1 200mm length of medium-duty hookup wire or 100mm of light gauge figure-8 wire (for speaker)
1 12m length 0.25mm enamelled copper wire (L1)
1 400mm length of 0.7mm-diameter tinned copper wire (LED lead extensions)
1 400mm length of 1mm-diameter heatshrink tubing (LED1-LED4 leads)
1 10mm length of 20mm-diameter heatshrink tubing (L1)
7 PC stakes (TP, 3 × GND, TP1, TPS, 2 × L1)

Semiconductors

1 MC1496P (DIP) or MC1496D (SOIC) balanced modulator (IC1)
1 TL072CP (DIP) or TL072CD (SOIC) dual op amp (IC2)
1 TDA7052A (DIP) or TDA7052AT (SOIC) BTL amplifier (IC3)
1 7809 3-terminal regulator (REG1)
4 2N5485 N-channel JFETs (preferably from the same manufacturer and batch) (Q1-Q4) (check eBay)
4 3mm blue LEDs (diffused lenses preferable) (LED1-LED4)
1 3mm red or green LED (LED5)
1 W04 bridge rectifier (BR1)
1 1N4148 signal diode (D1)
1 1N5819 Schottky diode (D2)

Capacitors

1 470µF 25V PC electrolytic
4 10µF 16V PC electrolytic
2 470nF MKT
12 100nF MKT
2 22nF MKT
2 1nF MKT
2 330pF NP0 ceramic
1 100pF NP0 ceramic
2 68pF NP0 ceramic
1 2-10pF trimmer capacitor (VC1)

Resistors (0.25W, 1%)

9 100kΩ	2 680Ω
1 47kΩ	2 560Ω
1 10kΩ	1 220Ω
1 6.8kΩ	1 150Ω
2 2.2kΩ	2 100Ω
1 1.2kΩ	1 82Ω
6 1kΩ	1 39Ω
1 820Ω	

Volume Control Board

1 single-sided PCB, available from the *EPE PCB Service*, code 23108142, 61 × 47mm
1 UB5 translucent blue plastic utility box, 83 × 54 × 31mm
1 Sharp GP2Y0A41SK0F 40-300mm distance measuring sensor (SENSOR1) (RS Components Cat 666-6568P, Littlebird Electronics DF-SEN-0143, Digi-Key 425-2819-ND, Element 14 code 1618431)
1 3-way PCB-mount screw terminal block, 5.08mm spacing (CON5)
1 3-pin header with 2.5mm spacing (for Sharp sensor)
1 M3 × 6mm machine screw
2 M3 × 10mm machine screws
3 M3 nuts
4 3mm ID washers
1 50mm length of 1mm clear heatshrink tubing (central wire between CON2,CON5)
1 300mm length of 1mm straight steel or aluminium wire (between CON2 and CON5)
1 120mm length of 6mm diameter heatshrink tubing (packing inside aluminium tubing)

Semiconductors

1 7805 3-terminal regulator (REG2)
1 1N4004 1A diode (D3)
2 3mm blue LEDs (diffused lenses preferable) (LED6,LED7)

Capacitors

1 1000µF 25V PC electrolytic
1 10µF 16V PC electrolytic

Resistors

2 470Ω 0.25W 1%

Extra hardware

1 800mm length of 10mm-diameter (OD) × 1mm-thick aluminium tubing (cut for 450mm antenna, volume control attachment and tripod stand)
1 350mm length of M5 or 3/16-inch zinc-plated threaded steel rod (cut to 2 × 62mm and 3 × 75mm)
10 M5 or 3/16-inch nuts to suit threaded rod (eg, nylon lock nuts in preference to half nuts)
1 151 × 90 × 19mm pine timber
1 29mm-OD frosted half-hemisphere hollow plastic ball (cut from ball salvaged from roll-on deodorant) (optional)

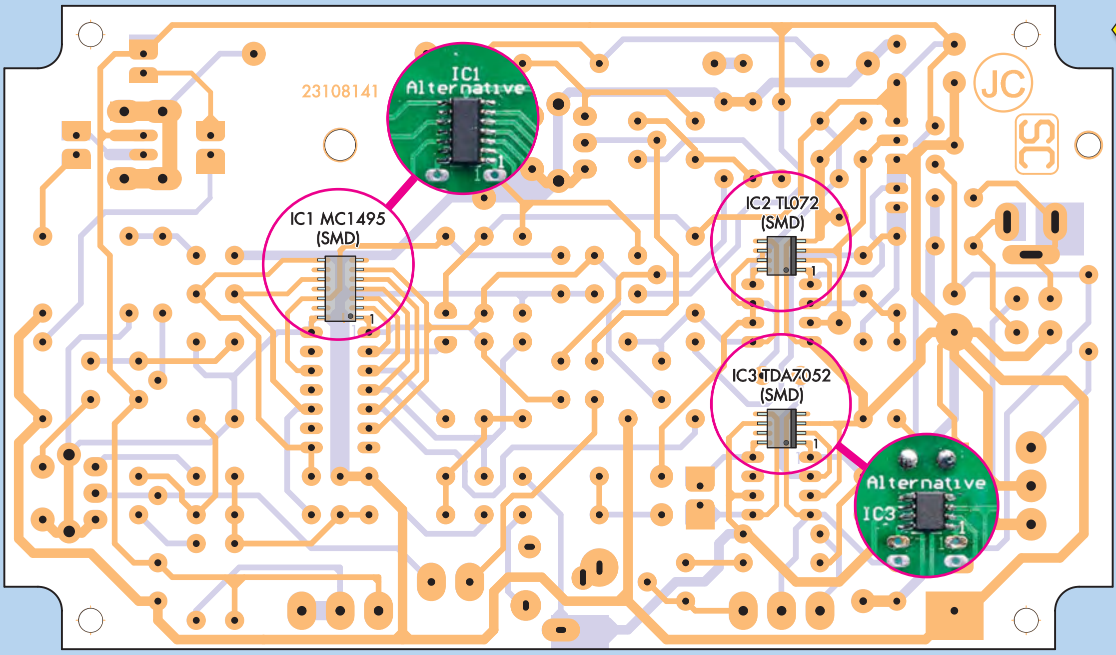

Fig.5: here's how to mount the alternative SMD ICs on the back of the PCB. Our prototype used SMDs for IC1 and IC3, as shown in the inset photos.

If you're not using SOIC ICs, or once you've finished fitting them, install the resistors. Be sure to push them all the way down so that they sit flush against the PCB before soldering their leads. Table 1 shows the resistor colour codes, but you should also check each one using a DMM before soldering it in position.

Next, fit any DIP ICs, either by soldering them directly to the PCB or using IC sockets. That done, fit PC stakes to the three GND positions (ie, TP GND and the GND pads adjacent to VR1 and T1), then TP1, TPS (adjacent to IC2) and for the two wiring points for coil L1. The two 3-way pin headers for LK1 and LK2 can then go in.

Diodes D1 and D2 are next on the list, taking care to ensure that they are correctly oriented. Bridge rectifier BR1 can also be installed at this stage, with its '+' pin positioned as shown.

Follow with JFETs Q1-Q4 and trimpots VR3 and VR4. Note that VR4 is oriented with its adjustment screw adjacent to LED5. The capacitors can then all go in, but be sure to orient the electrolytic types correctly. Table 2 shows the codes used on the low-value capacitors.

LED3 is installed next and must be pushed all the way down onto the PCB before being soldered. Its anode (A) lead is oriented as shown.

Once it's in, the two adjacent M205 fuse clips (used to connect the antenna) can go in. These must have their end-stop tabs broken off before installation, by bending them back and forth using small pliers.

These fuse clips are both mounted slightly proud of the PCB and their pins soldered on both sides of the board, to make a secure mounting receptacle for the antenna. **Do not push the fuse clips all the way down onto the PCB, as they could short to LED3's pads.**

The two oscillator coils, T1 and T2, can now be installed. These are both white-cored IF transformers and only go in one way, since they have three pins on one side and two on the other. Push them all the way down onto the PCB before soldering their pins and don't forget to solder the mounting pins on either side of the metal cans.

Once these parts are in, install switch S1, power socket CON1, 3-way screw terminal block CON2 and 3.5mm jack socket CON3. Note that the wire entry side for CON2 must go towards the adjacent edge of the PCB.

9V regulator

Regulator REG1 mounts horizontally, with its leads bent by 90° so that they go through the PCB holes. Secure its tab to the PCB using an M3 × 6mm screw and M3 nut before soldering the leads. Don't solder the leads first; if you do, the PCB tracks could crack as the screw is tightened.

Next, cut the shafts of VR1 and VR2 to suit the knobs that will be used and clean up the ends with a file. That done, snap off the small lug next to the threaded shaft bushing on each pot and install the two pots on the PCB.

The metal body of each potentiometer must be earthed to the PCB via a GND PC stake. For VR1, the GND stake is immediately adjacent and the pot's metal body is connected to it using a short length of tinned copper wire. Note that it will be necessary to scrape or file away a small section of the passivation layer on the pot's body to allow the solder to adhere.

By contrast, VR2's GND stake is some distance away, to the left of coil L1. It should be connected using medium-duty hookup wire. This earth position was necessary to remove background hiss from the *Opto-Theremin*'s audio outputs.

Front-panel LEDs

The remaining LEDs (LED1, LED2, LED4 and LED5) must all be mounted on 35mm lead lengths, so that they later protrude through the lid of the box. This means that you will have to extend their leads using short lengths of tinned copper wire.

Keep the anode leads slightly longer than the cathode leads, to make it easy to check the polarity when the LEDs are installed. It will be necessary to sleeve at least one lead of each LED

Table 1: Resistor Colour Codes

❏	No.	Value	4-Band Code (1%)	5-Band Code (1%)
❏	9	100kΩ	brown black yellow brown	brown black black orange brown
❏	1	47kΩ	yellow violet orange brown	yellow violet black red brown
❏	1	10kΩ	brown black orange brown	brown black black red brown
❏	1	6.8kΩ	blue grey red brown	blue grey black brown brown
❏	2	2.2kΩ	red red red brown	red red black brown brown
❏	1	1.2kΩ	brown red red brown	brown red black brown brown
❏	6	1kΩ	brown black red brown	brown black black brown brown
❏	1	820Ω	grey red brown brown	grey red black black brown
❏	2	680Ω	blue grey brown brown	blue grey black black brown
❏	2	560Ω	green blue brown brown	green blue black black brown
❏	1	220Ω	red red brown brown	red red black black brown
❏	1	150Ω	brown green brown brown	brown green black black brown
❏	2	100Ω	brown black brown brown	brown black black black brown
❏	1	82Ω	grey red black brown	grey red black gold brown
❏	1	39Ω	orange white black brown	orange white black gold brown

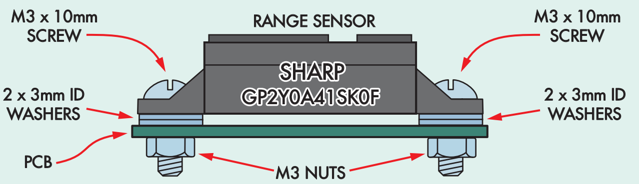

Fig.6: this diagram shows the mounting details for the Sharp optical distance sensor. Note the 3mm stacked washers used as spacers.

Table 2: Capacitor Codes

Value	µF Value	IEC Code	EIA Code
470nF	0.47µF	470n	474
100nF	0.1µF	100n	104
22nF	0.022µF	22n	223
1nF	0.001µF	1n	102
330pF	NA	330p	331
100pF	NA	100p	101
68pF	NA	68p	68

with heatshrink tubing after attaching the wire extensions – to prevent shorts.

Once the extensions are in place, mount the LEDs on the PCB (red for LED5, blue for the others), taking care to ensure that they are oriented correctly. It's a good idea to slide a 35mm-wide strip of cardboard between the leads of each LED when mounting it in position. It's then just a matter of pushing it down onto this spacer before soldering its leads.

Equalising coil

Equalising coil L1 consists of a bobbin and two ferrite core halves. The first step is to 'jumble-wind' (ie, randomly wind) 260 turns of 0.25mm enamelled copper wire onto the bobbin. When the winding is complete, lightly twist the two free ends together for about 2mm to prevent the winding from unravelling, then cut the leads to 20mm and scrape away the insulation from each end.

Next, cover the winding with a layer of insulation tape. Alternatively, shrink some 20mm-diameter heatshrink tubing around the bobbin. The coil can now be assembled onto the PCB, as follows:

1) Position one ferrite core section on the PCB and fit the bobbin in place.
2) Slide two 4mm-ID nylon or polycarbonate washers inside the bobbin, so that they rest on top of the inner part of the bottom core (these are needed to provide a 2.5mm spacing between the two cores).
3) Place the top core in position and secure the entire assembly to the PCB using an M4 × 25mm nylon or polycarbonate screw and an M4 nut. Be sure to orient the coil as shown on the parts layout diagram (Fig.4).
4) Solder the coil wires to the adjacent PC stakes.

Volume control PCB

That completes the main PCB assembly – now for the volume control board. Start by installing the two 470Ω resistors and diode D3, then fit regulator REG2. As before, be sure to secure the regulator's tab to the PCB using an M3 × 6mm screw and M3 nut before soldering the leads

Next, fit the 10µF and 1000µF electrolytic capacitors. As shown, the latter must be installed with its body lying horizontally and its leads bent down through 90° to go through its solder pads. The 10µF capacitor will also need to be bent over slightly so that it later clears the case lid.

The two blue LEDs can go in next. These are mounted with their bodies close to the PCB and are bent slightly towards the 470Ω resistor, so they do not later directly shine into the player's eyes. If you are not using a translucent case, then the LEDs will need to be mounted about 10mm proud of the PCB, so they later protrude through the case.

The distance sensor is installed by first soldering a 3-way pin header to the pins of the right-angle 3-way connector on the underside of its PCB. This is clearly shown in one of the accompanying photos. The sensor is then mounted on the volume control PCB and secured using M3 × 10mm screws and nuts, with two stacked M3 washers serving as spacers on each side – see Fig.6.

Tighten the screws down firmly before soldering the 3-way pin header to the PCB.

That's all we have space for this month. Next month, we'll describe how the two boards are assembled into their boxes, give the final mechanical assembly details and detail the simple test and adjustment procedure.

Win a Microchip PIC24FJ256DA210 Development Board

EVERYDAY PRACTICAL ELECTRONICS is offering its readers the chance to win a Microchip PIC24FJ256DA210 Development Board. The PIC24FJ256DA210 Development Board (DM240312) is a low-cost and efficient development board used to evaluate the features and performance of the PIC24FJ256DA210 with integrated graphics, mTouch and USB. The development board comes with a graphics display Truly 3.2 240×320 board (AC164127-4) to complete the two-board setup. It has a Microchip display connector V1, and allows developers to match with any of the listed 3.2, 4.3 TFT display, or the graphics prototype board available by Microchip.

KEY FEATURES

• PIC24FJ256DA210 16-bit microcontroller • Capacitive touch pads and switches • Microchip display connector V1 • USB connectors (embedded host/device/OTG) • PICtail Plus modular expansion slot • RS-232 serial port and associated hardware • Debug connectors supporting MPLAB ICD-3, MPLAB REAL ICE and MPLAB PICkit-3 • Graphics display Truly 3.2 240×320 board

The development board also provides a complete interface to MPLAB ICD-3, MPLAB REAL ICE, and MPLAB PICkit-3 emulator and debugger.

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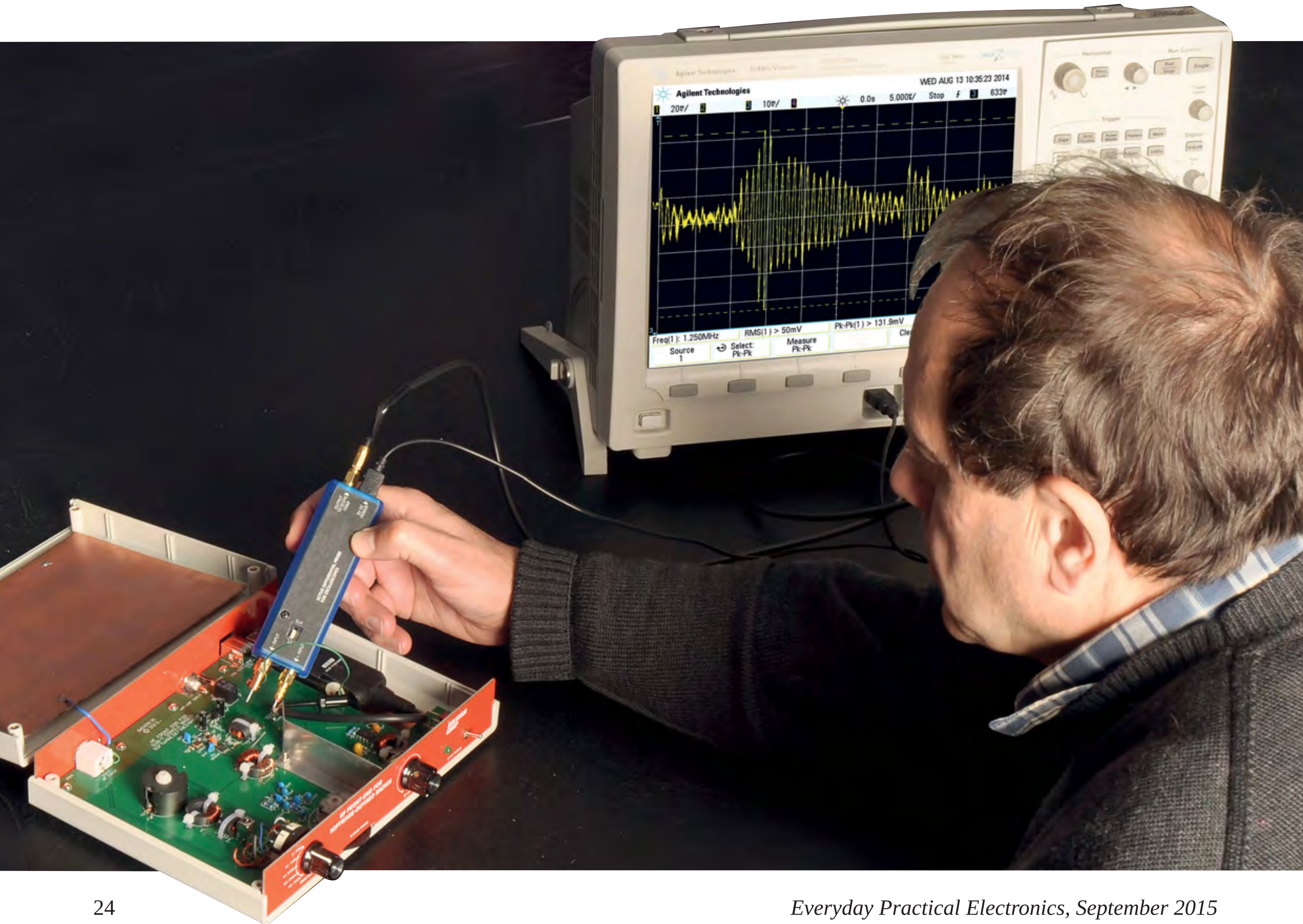
CLOSING DATE

The closing date for this offer is 30 September 2015.

WIDEBAND, ACTIVE DIFFER OSCILLOSCOPE

Using your oscilloscope to examine and measure high speed and high frequency circuits can be tricky if you use only the usual passive test probes supplied. Here's a design for a high performance, active differential probe which costs much less than commercially available active probes. It has very little circuit loading and a usable bandwidth of more 80MHz.



By JIM ROWE

ENTIAL PROBE

Do you know what is inside the 'passive' test probes supplied with most oscilloscopes, and how to use them to make reasonably accurate measurements at high frequencies? Oscilloscope probes can be complex and many factors can result in their performance falling away, especially at high frequencies. In addition they tend to disturb operation in the circuit being tested, making it difficult to make proper measurements.

It's because of the shortcomings of passive probes that some of the big manufacturers produce 'active' probes to provide a much higher input resistance together with a much lower input capacitance.

Originally, active probes used valves (vacuum tubes) at their input, but then when semiconductor technology came along, JFETs and MOSFETs made it possible to make active probes that were much smaller and easier to use.

It also became feasible to make 'differential' active probes, which overcame some of the remaining drawbacks with conventional 'single-ended' active probes – more about these shortly.

The big problem with commercial active probes is their price tags. Even the single-ended type can set you back well over £350, while the differential type can cost over £1000 apiece – more than most of us paid for our digital scopes!

In short, the only way that most of our readers are likely to be able to use an active probe with their scope is to build one.

This new active, differential scope probe design takes advantage of modern surface-mount components to deliver a high level of performance and it fits inside a compact case. Best of all, it can be built for much less than the cost of any currently available commercial active probes.

We estimate that you should be able to buy all of the components and build it for about a quarter of what you'll pay for the cheapest commercial active probe presently available.

Why differential?

Before we start describing the new probe and how it works, perhaps we should look at why a differential active probe tends to be better than a single-ended one.

A single-ended active probe is certainly a big improvement over most passive probes, offering high input resistance combined with very low input capacitance.

It tends to cause lower disturbance to the circuit under test, particularly at high frequencies – where the higher input capacitance of a passive probe causes increased circuit loading.

The high-frequency and transient response of the probe-plus-scope combination also tends to be better and smoother, due to better compensation and fewer reflections in the cable between the probe output and the scope input.

But there can still be problems when you're making HF measurements with a single-ended active probe. These problems are mainly associated with the 'ground clip lead', which is used to make the connection between the probe's input and the earthy or 'cold' side of the circuit under test.

As you can see from Fig.1A, even when the ground clip lead is quite short, it can introduce enough inductance (L_g) to reduce the effective signal voltage appearing at the actual input of the probe at high frequencies.

So the frequency response of the probe tends to droop at high frequencies, reducing the measurement reliability.

Also, the ground lead inductance can interact with the probe's input capacitance (C_{in}), resulting in resonances at specific high frequencies.

This can not only result in the probe producing unwanted loading on the circuit being measured, but can also produce spurious 'peaks and dips' in the measurements.

It is possible to minimise these problems by replacing the ground lead with a very short 'ground blade', providing a somewhat lower inductance than the usual 100mm-or-so long ground lead and clip.

Many of the commercial single-ended active probes come with this type of ground blade as an accessory. But a better solution is to change over to a differential probe, as shown in simplified form in Fig.1B.

As shown, the differential probe has two tips and is designed to measure the signal difference between the two – rather than the signal between either probe tip and ground.

In fact, the ground lead (or blade) is really only used to tie the circuit under test's ground to that of the probe and scope, to keep the voltages being measured within the probe's common-mode input range.

This means that if there is no significant voltage difference between the two grounds, the ground lead or blade may be regarded as optional.

Inside the differential probe there are two virtually identical input buffer amplifiers (one from each tip), each of which feeds one input of a third amplifier, the differential amplifier. This is where one of the two signals is subtracted from the other to send only the 'difference' signal out to the scope input.

This subtraction cancels out any 'common-mode' signal present at both probe tips, leaving only the 'difference signal' – the signal between the positive and negative probe tips, which is what we are trying to look at and measure.

Since the common-mode signal is essentially equal to the voltage V_{COM} at the probe's ground terminal, this explains why any voltage difference developed across the ground lead or blade inductance L_G is no longer a problem. It's simply cancelled out.

Before we leave Fig.1, you may be wondering why we've shown the output of the differential probe as having an amplitude of 2Vsig. Won't this cause a calibration problem, by giving the probe a gain of 2?

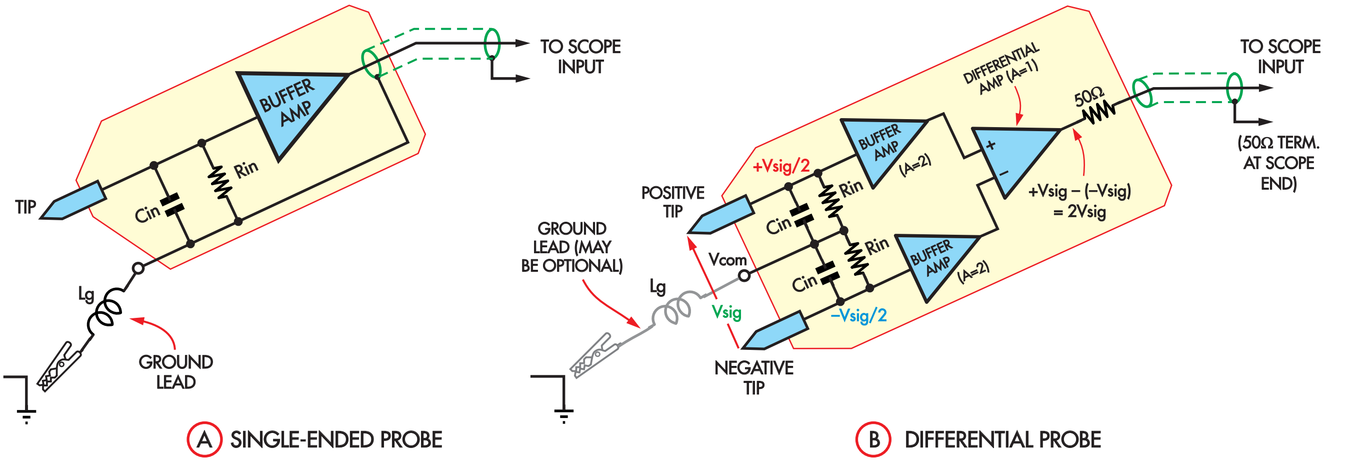


Fig.1: Comparing a 'single-ended' active probe (A) with a differential active probe (B). With a single-ended probe, the ground lead inductance Lg can cause problems at high frequencies, but a differential probe solves these problems.

Not really, because as shown in Fig.1B, there's a 'source termination resistor' of 50Ω fitted in series with the probe output. This is to match the characteristic impedance of the probe's output cable (normally 50Ω).

Then at the scope end of the same cable, another 50Ω shunt resistor is used to ensure that the cable is terminated correctly at that end too, to avoid reflections and consequent complications (like peaks and dips).

And the combined effect of the two termination resistors is to introduce an attenuation factor of 2:1 – bringing the overall signal gain of the probe and cable back to unity.

The probe's circuit

Now refer to Fig.3, which shows the complete schematic of our new probe. The two probe input tips plug into CON1 and CON2 at left, from where they each pass to the end contacts of S1a and S1b – the two sections of range switch S1.

Depending on the setting of S1, they each pass into the two input buffer amplifiers directly or via a series 9.0MΩ input divider resistor comprising three 3.0MΩ 1% resistors in series.

Then each signal passes through a DC blocking capacitance comprising a 1nF capacitor in parallel with a 10nF capacitor.

This combination has been chosen to give a lower input corner frequency of less than 30Hz, together with the smoothest possible upper frequency response.

Following the DC blocking capacitors the signals each pass through 27Ω

Specifications

An active differential probe for oscilloscopes, housed in a compact handheld case and operating from +5V DC, derived from any convenient source such as a USB port on a PC or digital oscilloscope. It provides tip area illumination via a white LED and a choice of two switched gain settings: 1:1 or 10:1.

Input coupling:	AC
Input resistance, each probe tip:	1MΩ nominal on the 1:1 range (1.0023MΩ); 10MΩ nominal on the 10:1 range
Input capacitance, each probe input socket to ground:	3.15pF approximately. (So capacitance tip-to-tip is approximately 1.6pF.)
Maximum DC voltage at probe tips:	±45V, both ranges
Maximum AC voltage input before overload, both probe tips:	2.0V peak-to-peak (700mV RMS) on the 1:1 range, 20.0V peak-to-peak (7.0V RMS) on the 10:1 range
Output impedance:	50Ω (Needs an output cable of 50Ω characteristic impedance, terminated in 50Ω at the scope end.)
Bandwidth (probe + output cable and termination):	25Hz – 80MHz +0.2dB/-3dB, both ranges; 60Hz – 50MHz +0.2dB/-0.5dB, both ranges; 150Hz – 40MHz +0.2dB/-0.3dB, both ranges
Overall transmission gain/loss:	On 1:1 range, 0.0dB ±0.6dB; On 10:1 range, -20dB ±1.0dB
Current drain from 5V DC supply:	Less than 40mA

This photo, close to life size, shows the *Active Differential Probe* in its handheld instrument case. It's a comfortable fit in the hand while applying the probe to the circuit under test.

overload-protection resistors, before reaching the gates of input buffer transistors Q1 and Q2.

These are BSS83 N-channel MOSFETs, designed especially for operating from a 5V supply voltage. We're using them as near-unity gain wideband source followers, to give high input impedance combined with the lowest-possible input capacitance.

The gates of both Q1 and Q2 are biased to +4.3V via the 1MΩ resistors. This bias level is chosen to provide a 'half-supply voltage' (+2.5V) level at the sources, which are direct coupled to the following ICs. The bias voltage is derived via the 2.7kΩ + 15kΩ voltage divider, with a 10µF bypass capacitor to provide filtering.

10:1 attenuator

The 1MΩ gate biasing resistors also provide the main component of input resistance for both input channels, when switch S1 is in the '1:1' position. Then, when S1 is moved to the '10:1' position, they form the lower elements in the 10:1 input dividers (in conjunction with the 9.0MΩ series resistors).

After passing through input buffer transistors Q1 and Q2, the two input signals pass through amplifiers IC1 and IC2.

These are AD8038ARZ wideband amplifiers, specified for operation from a single 5V supply and with a bandwidth of better than 150MHz (for a gain of 2.0).

Incidentally, we also looked at several other devices, including the AD818, MAX4414ESA and OPA356, but none performed as well as the AD8038ARZ. So the 8038 it is!

We are using them here as buffer amplifiers with a gain of 2.3, to compensate for the small loss in the input source followers Q1 and Q2 – while also providing the current drive capability to feed the inputs of difference amplifier IC3. The two 47µF capacitors connecting the 1.0kΩ 'lower feedback' resistors to probe earth are to maintain the low-frequency response.

IC3 is also an AD8038ARZ device, configured so that the positive-tip input signal is fed to its positive input (pin 3) while the negative-tip signal is fed to the negative input (pin 2).

The four 1kΩ resistors and 47µF + 10nF bypass capacitors ensure that IC3 does perform the desired subtraction of the two signals, so a '2Vsig' difference signal appears at its output (pin 6).

The two paralleled 100Ω resistors at the output of IC3 provide the 50Ω 'source termination' for the cable connecting the probe's output at CON3 to the scope input and the paralleled

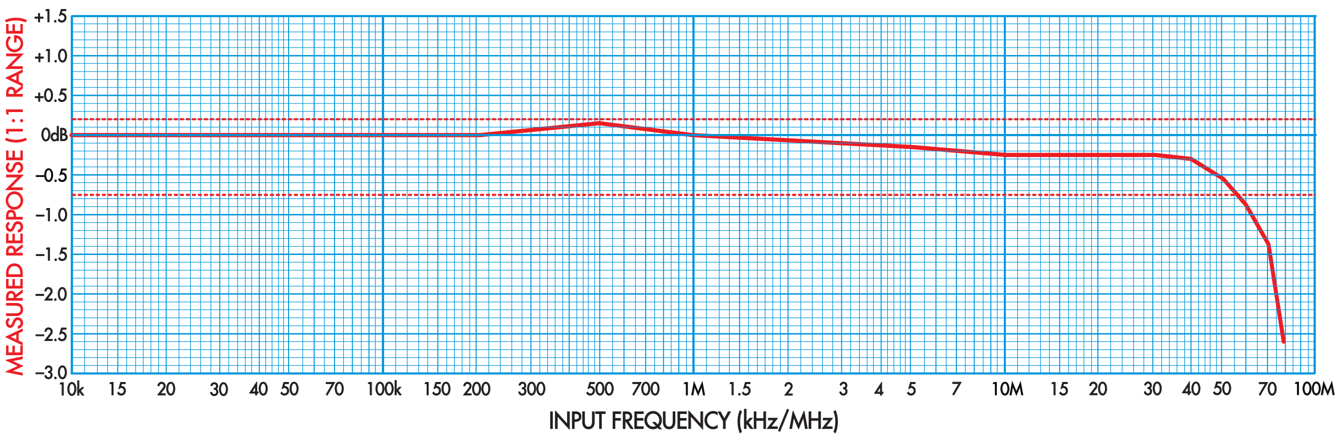
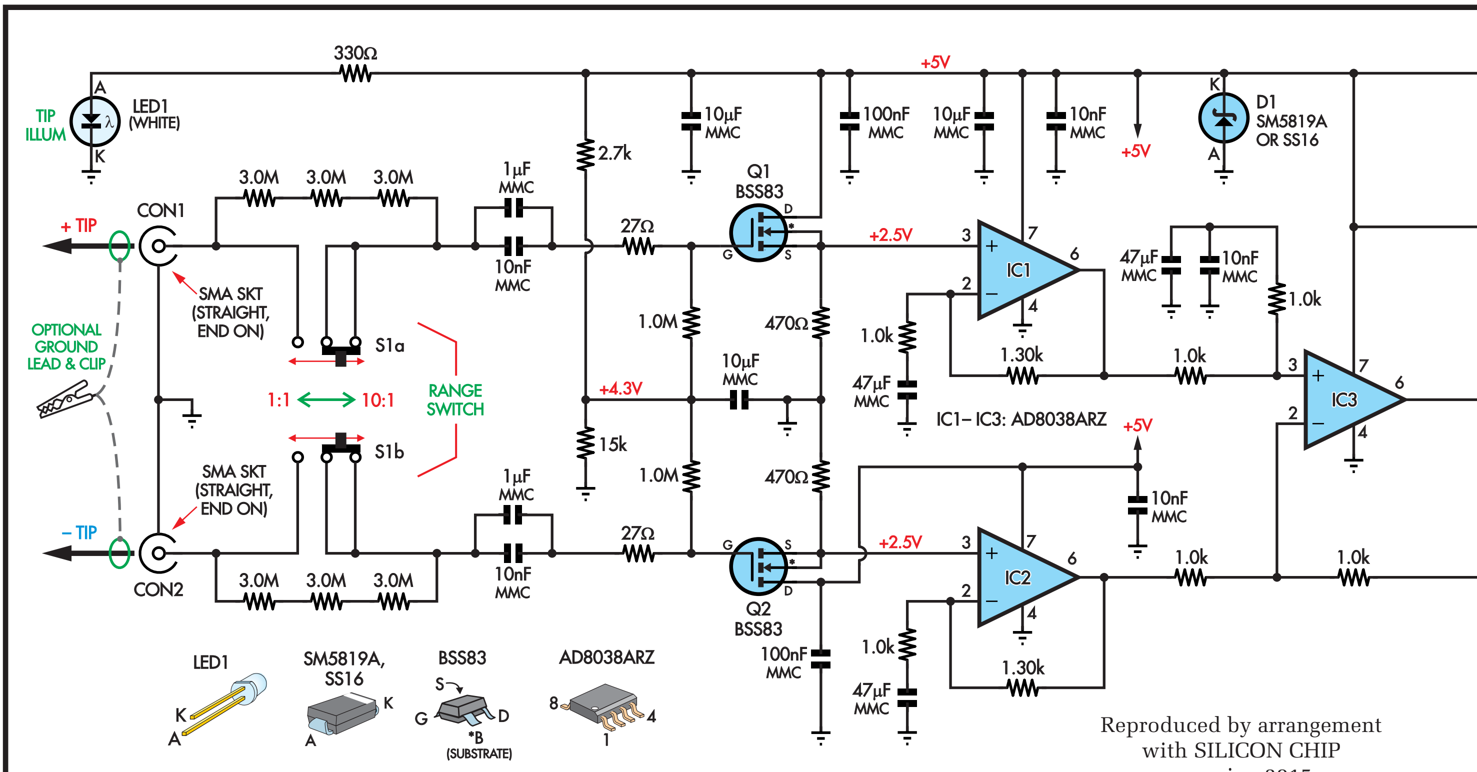


Fig.2: The upper frequency response of the differential probe, as measured on the 1:1 range. At the LF end it rolls off quite smoothly below 150Hz, with the –3dB point at around 25Hz.



ACTIVE DIFFERENTIAL RF SCOPE PROBE

100µF and 100nF capacitors provide DC blocking.

LED tip illumination

Finally, LED1, located at upper left is included to illuminate the area right in front of the probe's tips, to make connections easier.

Many of the up-market commercial active probes also provide this 'tip illumination', because when you are making measurements in high frequency circuits you'll almost certainly be using very short tips on the probe itself. This means that the probe body will not only shield the immediate area of the circuit being tested from a light source, but will also tend to block your view as well.

In other words, it's a very worthwhile feature and one which was easily provided at low cost.

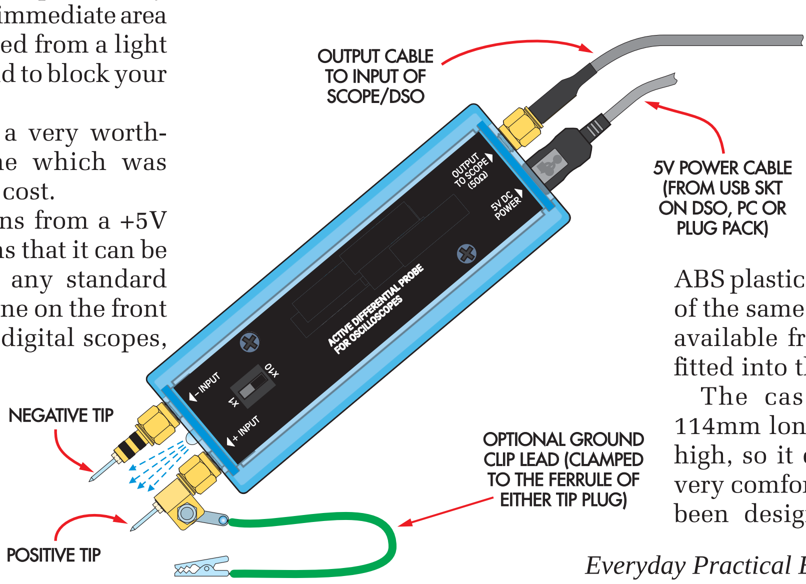
The whole probe runs from a +5V DC supply which means that it can be powered via virtually any standard USB port, such as the one on the front of many recent-model digital scopes, a USB port on your PC – or if neither of these are available, one of those low-cost 'USB charger/power pack' devices you can pick up for less than a tenner (preferably not a dodgy cheapo one!).

Since the total drain of the probe is less than 40mA, this should be well within the capability of most USB ports on DSOs and PCs.

CON4 is used to bring the +5V DC power into the probe. This is a USB micro type-B socket, which allows you to use a standard 'USB type A-plug to USB micro type-B plug' cable (as used to hook up tablet PCs and mobile phones to a PC or charger) to provide the probe with power.

100µH inductor L1 is used to filter the +5V input and remove any noise from the USB port or charger, while fuse F1 and diode D1 are used to protect against reversed-polarity damage.

These components do nothing if the 5V supply is connected with the correct polarity, but if the polarity should be reversed for any reason, D1 will immediately conduct and cause F1 to 'blow' – protecting both the probe circuitry and the 5V source from significant damage.



Construction

All of the probe circuitry and components are fitted onto a PCB measuring 103 × 26mm (code 04107141), which is available from the *EPE PCB Service*. This is designed to fit inside one half of a small handheld ABS plastic case, with a screening PCB of the same size (code 04107142, also available from the *EPE PCB Service*) fitted into the other half of the case.

The case itself measures only 114mm long, 36mm wide and 24mm high, so it can be held in your hand very comfortably. In fact, the case has been designed to house hand-held

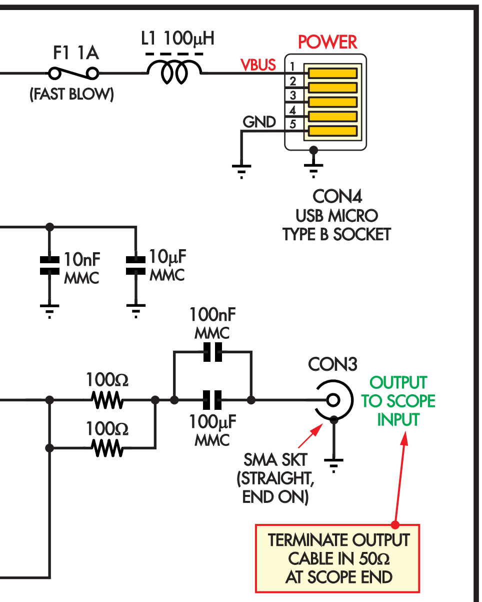


Fig.3: The probe's full circuit schematic. All components except range switch S1 and LED1 are SMD devices.

equipment such as this. It comes from Hammond Manufacturing.

The small SMA sockets (CON1 and CON2) used for connection of the probe's input tips are mounted at one end of the case, along with the white LED1, which illuminates the tip. Two sockets are mounted at the other end, SMA output socket (CON3) along with CON4, the USB micro B socket for the probe's 5V DC power.

All of the components used in the probe are mounted directly on the main PCB, and all but two of the components are SMDs (surface-mount-devices).

Parts List – Active Differential Oscilloscope Probe

1	ABS instrument case, 114 × 36 × 24mm	(Hammond 1593DTBU (see their advert below for contact details))
2	PCBs, 103 × 26mm, available from the *EPE PCB Service*, coded 04107141 & 04107142	
1	100µH SMD inductor, 1.6A rating (L1)	(Murata 48101SC)
1	1A SMD fuse, 0603 fast acting (F1)	(Cooper Bussman 0603FA1-R)
1	DPDT/DIL slide switch, raised actuator (S1)	(TE Connectivity ASE 2204)
3	SMA socket, end launch, PCB edge mtg (CON1,2,3)	(Emerson Connectivity 142-0701-806 or Multicomp 19-70-4-TGG)
1	Micro USB type B socket, SMD (CON4)	(FCI 10103594-0001LF or Molex 105017-0001)
8	Self-tapping screws, 6G × 5mm long	

Semiconductors

3	AD8038ARZ SOIC8 video amplifier (IC1,2,3)	(RS Components order code 523-6872)
2	BSS83 MOSFETs, SOT-143 SMD pkg (Q1,2)	(element14 order code 108-1312)
1	3mm white waterclear LED (LED1)	
1	60V 1A Schottky diode, DO214AC SMD pkg (D1)	(SS16 or SM5819A)

Capacitors

1	100µF MLCC, SMD 1210, X5R dielectric 6.3V rating	(Code 107)	NB: not
3	47µF MLCC, SMD 1210, X5R dielectric 6.3V rating	(Code 476)	all SMD
4	10µF MLCC, SMD 1210, X7R dielectric 16V rating	(Code 106)	capacitors
2	1µF MLCC, SMD 1206, X7R dielectric, 50V rating	(Code 105)	are marked.
3	100nF MLCC, SMD 1206, X7R dielectric 50V rating	(Code 104)	If in doubt,
6	10nF MLCC, SMD 1206, X7R dielectric 50V rating	(Code 103)	measure!

Resistors (all 0.125W 1%, SMD 1206)

6	3.0MΩ		(Code 3M0 or 3004)
2	1.0MΩ		(Code 1M0 or 1004)
1	15kΩ	The codes shown here	(Code 15K or 1502)
1	2.7kΩ	are the two most common	(Code 2K7 or 2701)
2	1.30kΩ	but there are others! If in	(Code 1K3 or 1301)
6	1.0kΩ	doubt, check all SMD	(Code 1K0 or 1001)
2	470Ω	resistors with your	(Code 471 or 470R)
1	330Ω	multimeter as you would	(Code 331 or 330R)
2	100Ω	any doubtful resistor.	(Code 101 or 100R)
2	27Ω		(Code 270 or 27R)

The two through-hole exceptions are slide switch S1 and LED1. Switch S1 is mounted under the PCB and LED1 is mounted above it with its leads bent forward by 90°, so that the LED's body can protrude through the 'front end' of the case between the two input sockets.

The component overlay diagram of Fig.4 shows the location of all components, together with their orientation. When assembling the PCB, use a fine-tipped soldering iron – preferably one with temperature control.

We suggest fitting the components to the PCB as follows: first fit USB micro socket CON4, taking great care when soldering its five very small contacts at the rear. Then mount the resistors and capacitors, followed by fuse F1 (which is very tiny). Then fit diode D1, MOSFETs Q1 and Q2 and the three ICs.

Next, fit the three SMA sockets (CON1, CON2 and CON3), which slide onto the front and rear edges of the PCB, with their centre pin resting on (and soldered to) the centre pad at the top of the PCB.

Their 'side prongs' solder to the matching pads on each side, on the top and bottom of the PCB.

Inductor L1 comes next, followed by LED1 on the top of the PCB and switch S1 underneath it in the position shown.

When you are fitting LED1 make sure you mount it vertically with the underside of its body about 13mm above the top of the PCB. After the leads are soldered they can both be bent forward (left) by 90°, so the LED can protrude from the centre hole in the case front end panel.

Finally, fit slider switch S1. This is in a 6-pin DIL package, which mounts under the PCB with its pins coming up through the matching holes. Make sure you push the switch body firmly against the underside of the PCB before you solder its pins to the pads on the top of the board.

Your PCB assembly should now be complete, with all that remains being to connect the shield PCB copper to the ground copper on the main PCB. This can be done using a short length of light hookup wire – baring a few millimetres at each end so that the ends can be soldered into the 'via' holes at the rear of each PCB, as shown in Fig.4.

Preparing the case

Now prepare the case. This involves drilling three 7mm holes in the removable 'front' end panel (for CON1, CON2

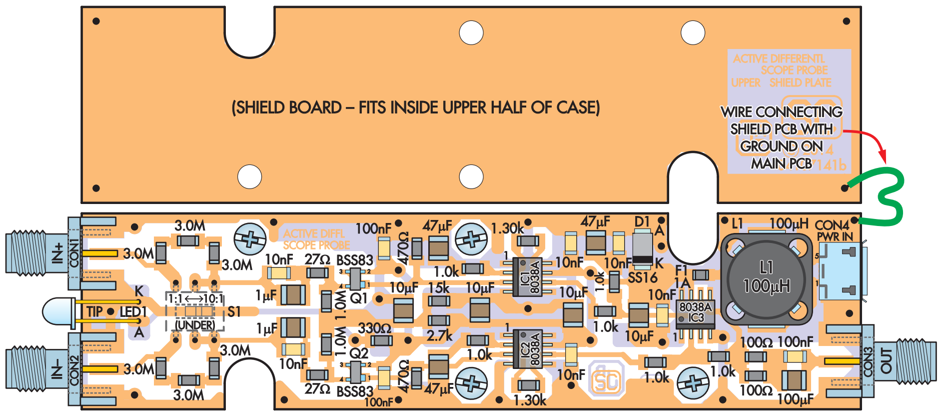
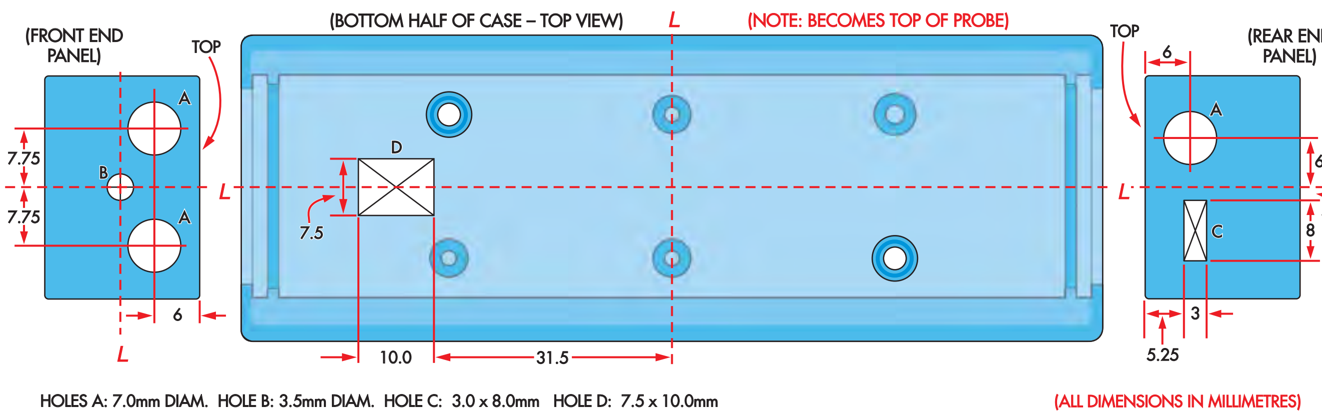
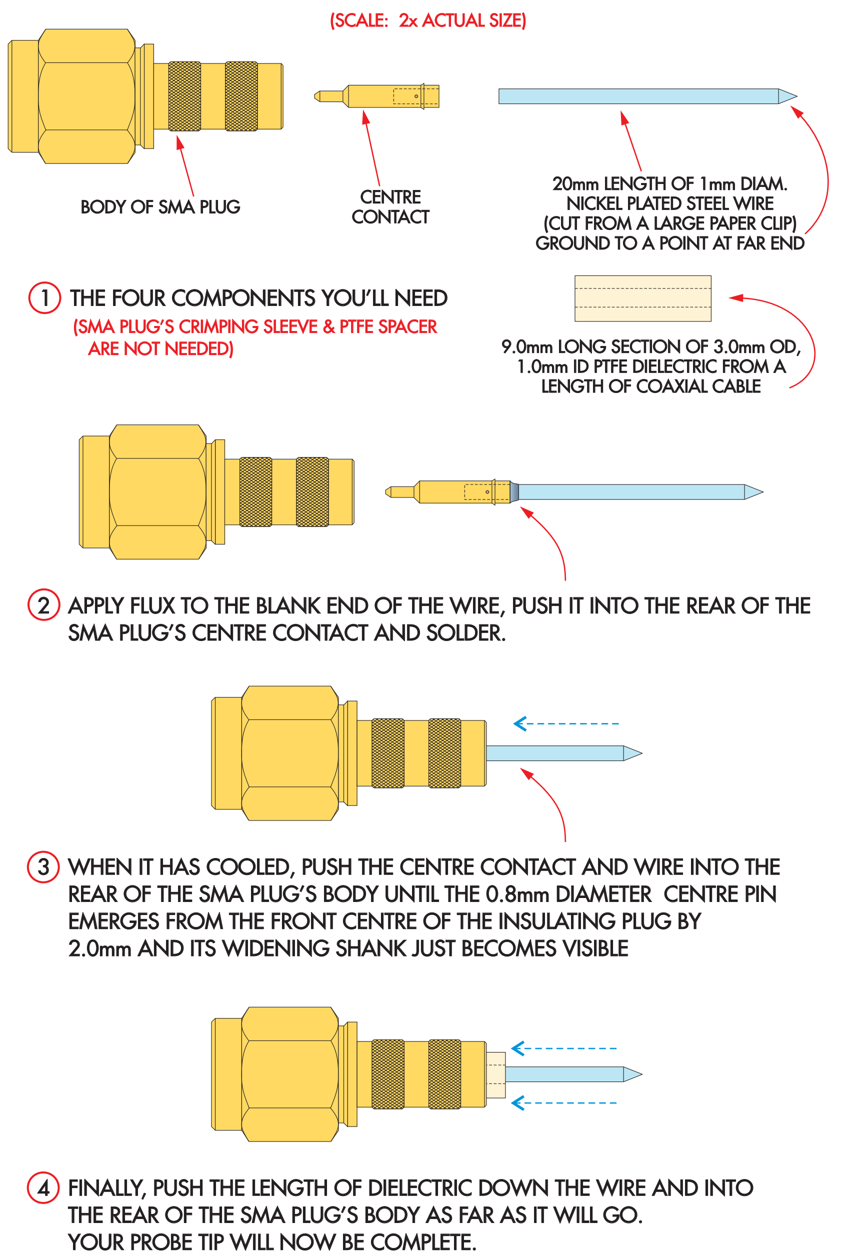


Fig.4 (above): the component overlay for the main PCB with the shield board (which contains no components) above. It is connected to the main board by the short link as shown. The main board fastens to the bottom of the case, which becomes the top, while the shield is secured to the top of the case, which becomes the bottom!
Below is a same-size photo of an early prototype main PCB, actually mounted in the case. Take no notice of the 'AD818' labelling – we actually used AD8038s – as shown on the PCB overlay.



Fig,5: drilling and cutout details of the Hammond Manufacturing 'Hand Held Instrument Case', shown 1:1. The only slightly difficult holes are the cutouts for the USB socket on the rear end panel and the switch on the lower half of the case.

Fig.6: Making a basic probe tip using an SMA plug.



and LED1), together with another round hole in the 'rear' end panel for CON3.

Then there's an 8 × 3mm rectangular hole to be cut in the rear end panel as well (for access to CON4), and finally a 10 × 7.5mm rectangular hole cut in the bottom half of the case (which becomes the top) for clearance around S1 and access to its actuator.

The location and size of all of these holes is shown in Fig.5. You might also want to make a 'dress' front panel, to give your probe a professional look and help in using it. Artwork for the front panel is also shown in Fig.7.

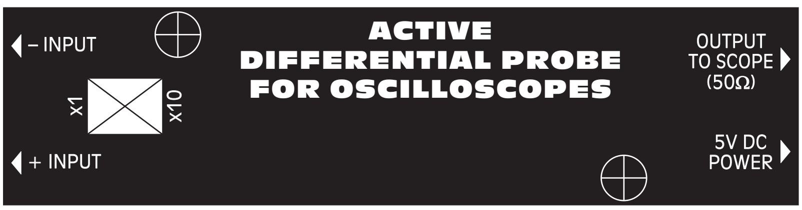


Fig.7: same-size front panel artwork to photocopy and glue to the hand-held instrument case for a professional finish.

The two ends of the case, with their drilling/cutouts to suit the three SMA sockets, USB socket and white LED.

You can make a photocopy of this, (or you can download it from **www.epemag.com** and print it), hot laminate it (or use self-adhesive book cover film) for protection and then attach it to the front panel using double-sided adhesive tape – after cutting it to size and also cutting out the clearance holes for the case assembly screws and S1.

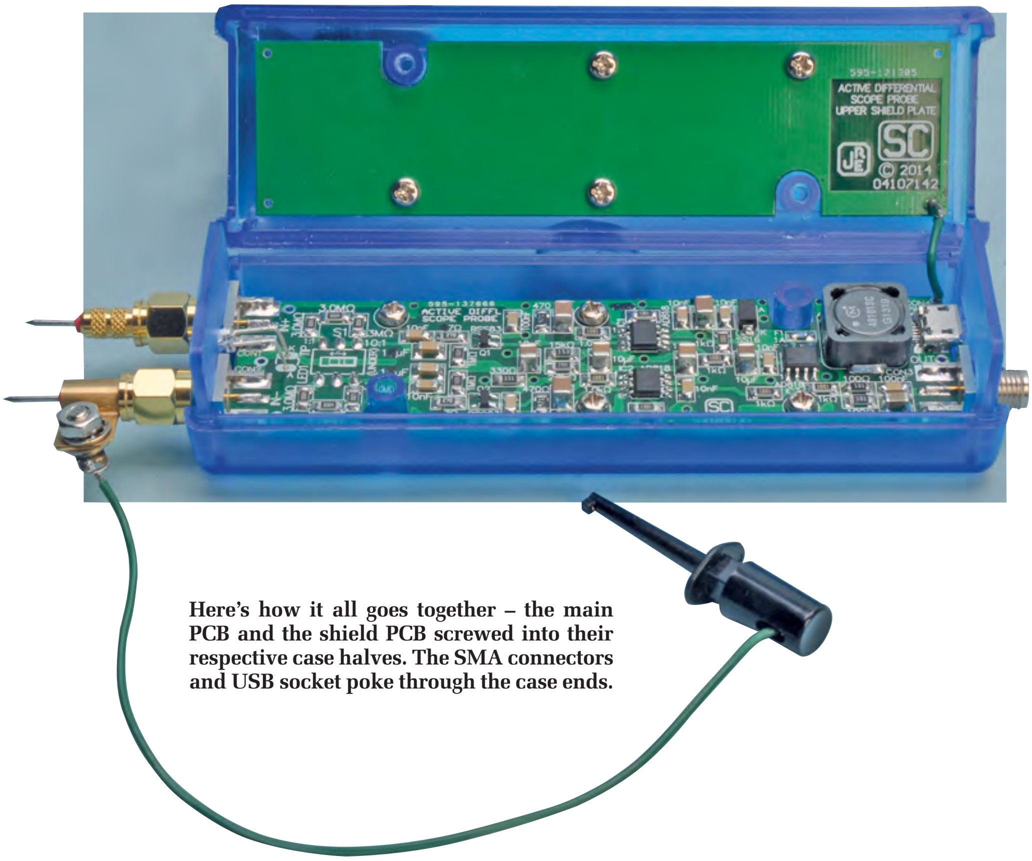
Assembly

Now slip the front end panel of the case over CON1, CON2 and LED1 at the 'front' end of the main PCB, and the rear end panel over CON3 at the rear end of the PCB.

Then lower the complete main PCB-plus-end-panels assembly down into the bottom half of the case (which becomes the top), with the two end panels passing into the moulded slots and S1 passing down through its matching slot.

Once this main board assembly is down as far as it will go, you can secure it firmly in position using four 5mm-long 6G self-tapping screws – mating with the holes in the moulded standoffs underneath.

Then the shield PCB can be fixed into the other half of the case, using another four of the same screws.



Here's how it all goes together – the main PCB and the shield PCB screwed into their respective case halves. The SMA connectors and USB socket poke through the case ends.

Other uses

A differential probe can also be handy for measuring signals which are relative to other voltages in a circuit. Both signals must be within the probe's common-mode input range and given that the probe is AC-coupled, you will only get the AC component of that signal.

For example, if you have a circuit with a signal that's relative to a 'half supply' rail, there may be ripple or signal injected into this rail. So, using the differential probe would allow you to see the signal with this unwanted component removed.

Many scopes can perform this function using 'math' mode, but that requires the use of two of your precious scope inputs and the result is generally a lot better when the subtraction is performed in the analogue domain.

With this method, the circuit ground can remain earthed, allowing easy simultaneous measurement of the signal.

The final assembly step is to invert the case half with the shield PCB and lower it down over the half with the main PCB, so that each end panel slips into the moulded slots as before.

Then upend it and fasten it all together using the two countersink-head self tappers supplied and your active differential probe should be complete.

Making the probe tips

The simplest way to make 'basic' probe tips for the project is probably to base them on an SMA male connector, as shown in Fig.6.

This is the way I made the probe tips you can see in the photos, basing them on an Amphenol Connex type 132113 SMA plug; only the plug body and the centre contact are used – the crimping sleeve and PTFE spacer are not needed.

The steps in making the tips are shown overleaf. The actual tips are 20mm lengths of 1mm diameter nickel plated steel wire, cut from a large paper clip.

You might like to make a second pair of tips, fashioned in the same way but with longer lengths of wire – say 30mm long – with a 'crank' in the centre to allow their tip spacing to be adjustable. This would be done simply by loosening their plug bodies and then rotating the tips as needed to set the tip spacing before tightening them again.

Ground clip lead

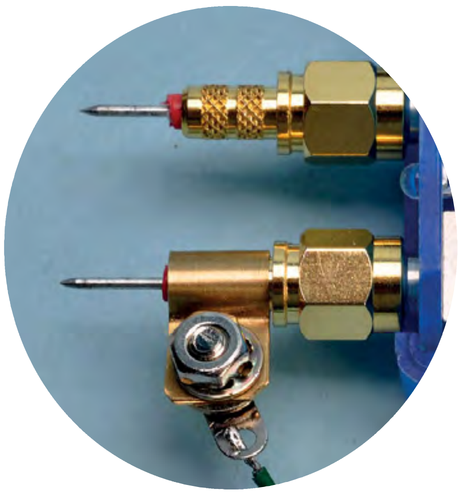
As mentioned earlier, a ground clip lead is often not necessary when you are using a differential probe of this kind. However, you might like to make one up, so it will be available in situations where you may need it – or at least to see if it has any effect.

An easy way to make a suitable clip lead is to connect a suitable clip to one end of a length of flexible insulated hookup wire and then fit a small solder lug to the other end.

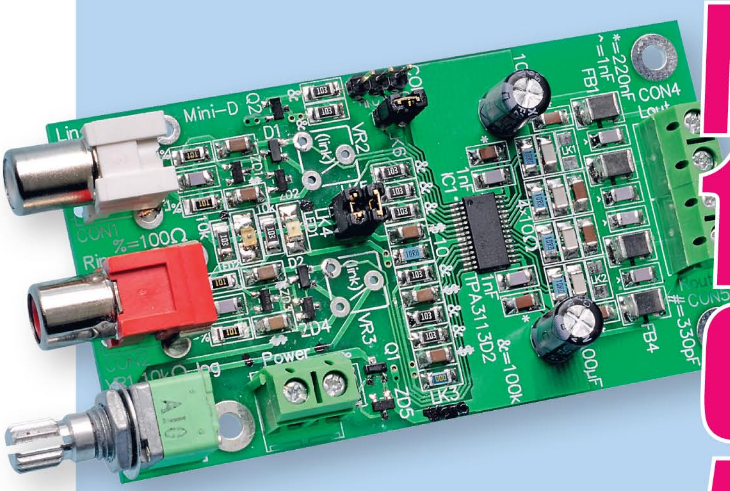
The solder lug can then be attached securely to a small clamp made of thin brass sheet and bent into a 'P' shape with an inner loop diameter of 4.5mm, so it will slip over the 'crimp ferrule' of one of your probe tip plugs (it doesn't matter which one).

Then a 3mm hole is drilled in the centre of the flat sections of the clamp, so a 6mm-long M3 screw and nut can be used to attach the solder lug of the ground lead, while at the same time fastening the clamp to the plug ferrule.

The close-up photograph above shows the idea. By the way, you don't have to make the ground clip lead particularly short, because its inductance is not critical when you are using a differential probe. So feel free to make it any convenient length.

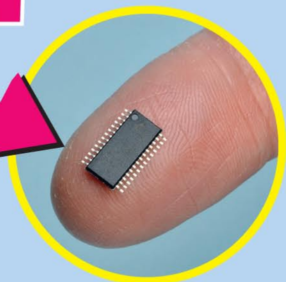


Close-up of ground clip construction



Mini-D Stereo 10W/Channel Class-D Audio Amplifier

This little chip can deliver a whopping 30 watts! With no heatsink!



Main features

- Stereo or mono Class-D amplifier on a single, small PCB
- No heatsink required
- Low EMI
- DC power supply, wide operating voltage range
- Drives one/two 4-8Ω speakers
- Selectable gain
- On-board volume control
- RCA input sockets
- Shutdown mode
- Output short-circuit protection
- DC offset protection
- Over-temperature shutdown with auto resume
- Selectable output power limit with soft clipping
- Low quiescent current
- Reversed supply polarity protection
- Input signal overload protection
- Power and fault indicator LEDs
- Under-voltage and over-voltage lock-out

This tiny class-D amplifier module can work in two modes. In stereo, it can deliver more than 10W per channel or you can connect its output channels in parallel to deliver more than 25W into a single speaker. It is up to 91% efficient, with selectable gain, volume control and other features such as a low-power shutdown mode and over-temperature, over-current, short-circuit and speaker protection.

By NICHOLAS VINEN

HOW CAN A CHIP this small deliver so much power? And how can it deliver so much power without needing a big heatsink? The answer to both questions is Class-D operation. It's a switching amplifier and its efficiency can be over 90%. High efficiency is also good if you want to run it from a battery since it will last longer. And if running from mains, you don't need a bulky power supply; a 1A plugpack should be more than adequate.

We published our first switching amplifier design, the *CLASSiC-D*, in November and December 2013. It's a powerful beast, able to deliver up to 250W into a 4Ω load, or 500W into an 8Ω load (bridged) with low distortion. Lots have been built since its publication.

But while you may want the high efficiency of Class-D, the *CLASSiC-D* is simply too big and expensive for many applications where you only need a few watts of audio, perhaps running off a small battery – for busking, for example. Or say you want to build a pair of self-powered computer

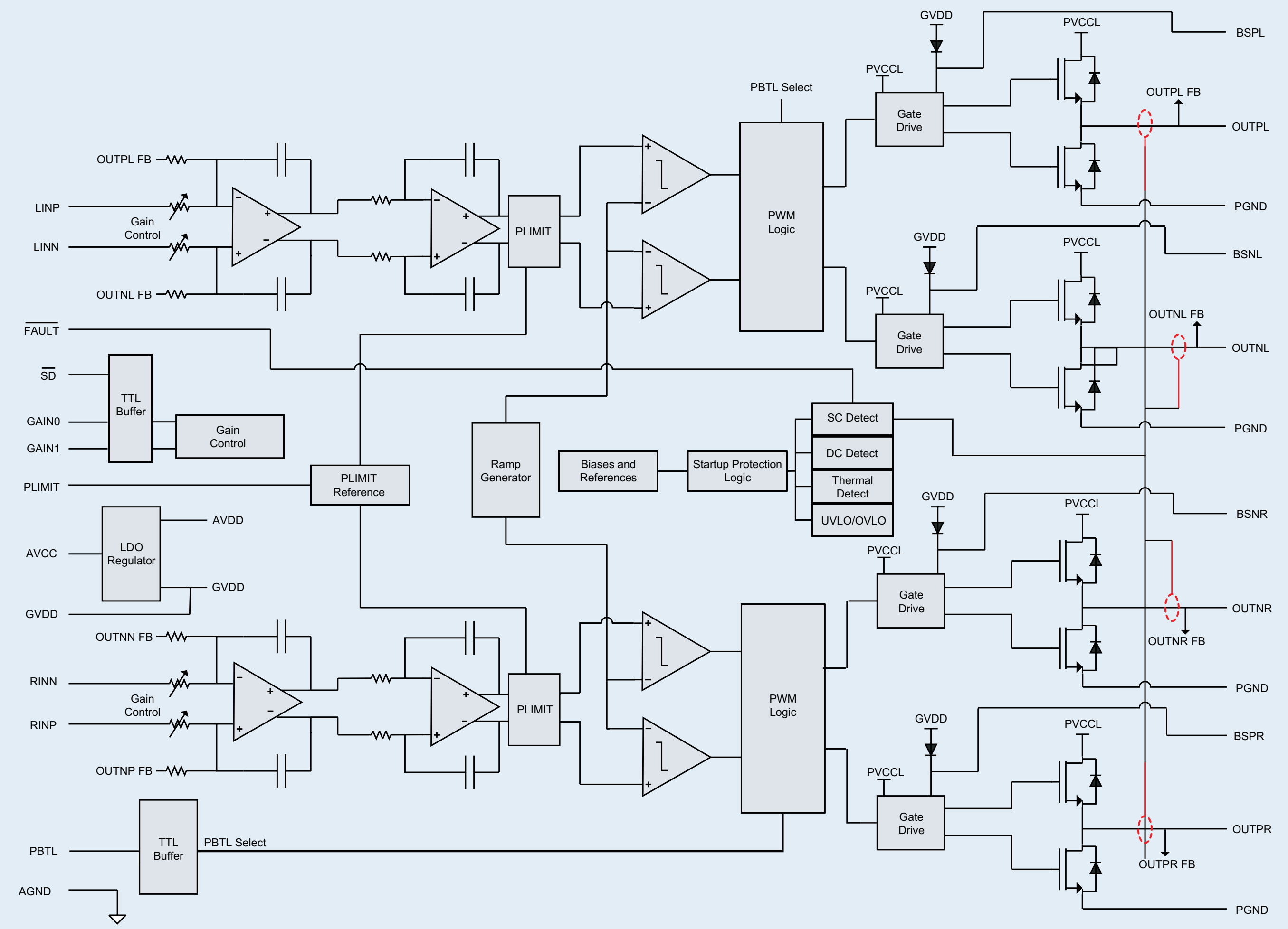


Fig.1: block diagram of the TPA3113D2 Class-D audio amplifier IC. The left and right channel differential inputs are buffered and fed to Schmitt trigger stages where they are compared against a ramp (triangle) signal. The resulting PWM signals are then fed to PWM logic blocks, which then drive two bridge-mode stereo switching amplifiers.

speakers. Whatever the reason, a few watts can go a long way.

That's where our *Mini-D Amplifier Module* comes into its own. It's based on the Texas Instruments TPA3113D2, which contains two complete bridge-mode stereo switching amplifiers. It's so efficient that it doesn't need a heatsink for normal program material; the PCB itself dissipates the heat.

Only a simple output filter is required to minimise the RF interference generated by its switch-mode operation. This consists of four ferrite beads and four ceramic capacitors – eight components for the two channels. All the components are surface-mount types, selected so that they are simple to solder to the PCB.

Because the *Mini-D Amplifier Module's* outputs are bridged, it has good power delivery, even with moderate supply rails. With a 12V supply, it can deliver at least 5W per channel into 8Ω speakers or 2 × 10W into 4Ω loads. More power is available with higher supply voltages.

Unusually, the *Mini-D Amplifier Module* can also operate in mono mode, with the outputs paralleled. This doubles its current capability, allowing more power into low-impedance loads, eg, 25W or more into 4Ω.

By the way, we've said this in the past, but it bears repeating: while the output transistors in class-D amplifiers spend most of their time either on or off, they aren't really 'digital' amplifiers. While there may be some digital circuitry involved, they still work on the principle of analogue negative feedback to generate the correct output waveform for a given input signal.

Class-D amplifier operation

We won't go into the full theory of how a class-D amplifier works, but let's look at the functional block diagram of the TPA3113D2 IC (Fig.1) which is the heart of the circuit.

The two inputs are differential. Looking at the left channel, the signals are fed to LINP (in-phase) and LINN (ground/out-of-phase) at top left. The feedbacks from the switching outputs, OUTPL FB (positive) and OUTNL FB (negative), pass through low-pass RC filters internal to the IC and these four signals all go into a differential amplifier, which performs this analogue computation:

(LINP – LINN) x GAIN – (OUTPL – OUTNL)

The GAIN setting is determined by the state of two digital inputs, GAIN0 and GAIN1, which control the resistances in this part of the circuit to select an effective gain of 20dB, 26dB, 32dB or 36dB. The output of this differential amplifier then passes through another RC low-pass filter, to further attenuate the switching artefacts in the signals, and then into a differential buffer.

During normal operation, with the output correctly tracking the input

Parts List

1 double-sided PCB, available from the *EPE PCB Service*, coded 01110141, 46 × 85mm
4 HI1812V101R-10 ferrite beads, SMD 4532/1812 (FB1-FB4) (element14 2292377)
2 PCB-mount switched RCA sockets, white and red (CON1-CON2) OR
2 2-way pin headers plus shielded cable, header plugs and chassis-mount RCA sockets
3 2-way mini terminal blocks, 5.08mm spacing (CON3-CON5)
1 3-way pin header, 2.54mm pitch (CON6)
3 shorting blocks
1 10kΩ dual-gang 9mm log potentiometer (VR1) OR
2 10kΩ mini horizontal trimpots (VR2-VR3) OR
1 20mm length tinned copper wire or two component lead off-cuts
3 2-way pin headers, 2.54mm pitch (LK4-LK6)
3 tapped spacers with M3 × 6mm machine screws (optional, for mounting)

Semiconductors
1 TPA3113D2PWP Class-D audio amplifier IC, HTSSOP-28 (element14 1762987)
1 IRFML8244 N-channel MOSFET, SOT-23 (Q1) (element14 1857298)
1 BSS84 P-channel MOSFET, SOT-23 (Q2) (element14 1431318)
5 5.6V Zener diodes, SOT-23 (ZD1-ZD5) (element14 1431238)
2 BAT54A dual Schottky diodes, SOT-23 (D1,D2) (element14 2114869)
1 high-brightness green LED, SMD 3216/1206 (LED1) (element14 2217905)*
1 high-brightness red LED, SMD 3216/1206 (LED2) (element14 1226389)*

Capacitors (all SMD 3216/1206 unless stated)**
2 100µF 25V low-ESR radial electrolytics
7 4.7µF 25V X7R ceramic (element14 1828835)
6 220nF 50V X7R ceramic (element14 1327724)
8 1nF 50V NP0/C0G ceramic (element14 2280692)
4 330pF 50V NP0/C0G ceramic (element14 3606090)

Resistors (all SMD 3216/1206 1%)**
9 100kΩ (element14 1811974)
2 10kΩ (element14 1811973)
2 100Ω (element14 1632521)
5 10Ω (element14 1591420)
2 4.7Ω (element14 2142059)
2 0Ω (element14 1632520) (LK1-3)

* or use 2 × 2-pin headers with off-board LEDs
** SMD 2012/0805 size parts can also be used

(after gain is taken into consideration), the output of these amplifiers will be virtually nil, ie, the two differential lines will be at the same potential. Any deviation from this state means that the amplifier output must swing one way or the other.

The buffered signal passes through the PLIMIT block, which allows an external voltage to limit the maximum output swing, for speaker overload protection. The signals then pass into a pair of Schmitt-trigger comparators, where they are compared against a ramp (triangle) signal, generated by an internal oscillator.

This is a common method for producing PWM (pulse-width modulation), typically used in motor control circuits. The main difference here is that the operating frequency is much higher; around 310kHz. This is necessary to allow accurate reproduction of audio signals up to 20kHz.

PWM output

These signals then pass through the PWM logic to the MOSFET gate drivers and then the totem-pole output stages, consisting of N-channel MOSFET pairs.

This chip uses a 'centre-aligned' or 'dual-ramp' PWM, a different modulation scheme to that used in many other class-D amplifiers. This is shown in Fig.2 and is possible because the TPA3113D2 always operates in bridged mode. In the quiescent condition, both outputs are driven in-phase with a 50% duty cycle (top of Fig.2) and this results in no current flowing in the speaker(s) or filter at all.

To drive the output positive, the duty cycle of the positive output is increased, while the negative-output duty cycle decreases (middle of Fig.2). This is done by shifting both the leading and trailing edges of both waveforms. Since none of these edges line up, this spreads RF emissions out, making them easier to filter. To drive the output negative, the reverse condition occurs (bottom of Fig.2).

Since the output transistors are N-channel MOSFETs, a supply above the positive rail is required for the upper gate drive. This is generated by four 220nF capacitors between the OUTPL and BSPL terminals, OUTNL and BSNL etc. When the respective output is low, its capacitor charges through an internal diode from GVDD (~7V) and when the output goes high, the capacitor charge maintains the associated boost pin 7V above that output, sufficient to keep the upper MOSFET conducting.

The block diagram also shows the protection circuitry, including short-circuit detection, output DC offset detection, high temperature detection and under/over-voltage lock-out. Should any of these fault conditions occur, the output drivers are all switched off. The over-temperature cut-out kicks in when the die temperature hits 150°C and operation resumes once it has dropped by around 15°C.

When the chip is running in mono mode, as set by the PBTL input pin, the PWM logic is modified slightly so that OUTPL and OUTNL carry an identical signal. At the same time, OUTPR and OUTNR are both driven with the same out-of-phase PWM signal, allowing the pairs of outputs to be paralleled.

Speaker wires

Because 'centre-aligned' PWM is used in this chip, only a very simple output filter is required to minimise the amount of RF interference generated. This consists of just four ferrite beads and four ceramic capacitors. The data sheet states that the ferrite bead output filter is sufficient for twisted speaker wires up to 1.2m long. We imagine that standard figure-8 speaker wires should also be OK, given that the conductors are in close proximity.

If you want to use longer speaker leads or are particularly concerned about radio interference, you can add

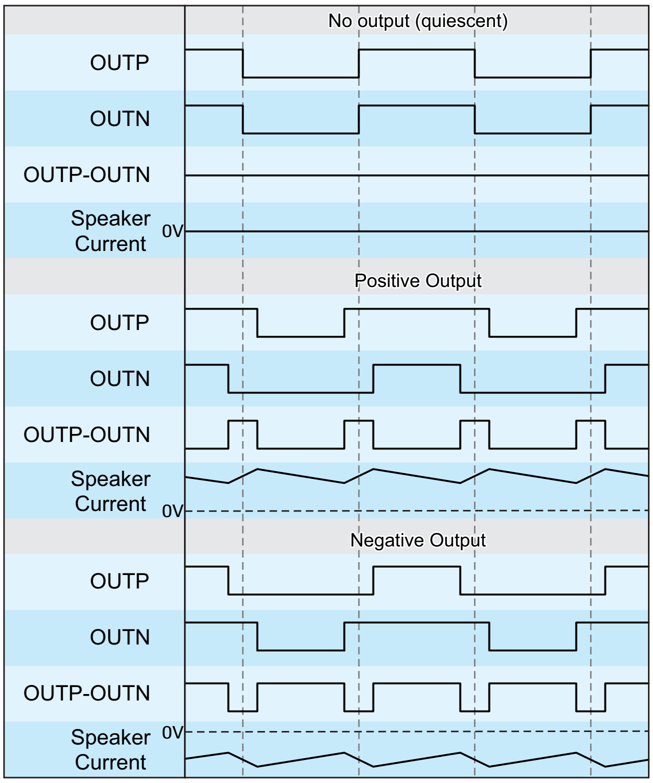


Fig.2: the quiescent (top), positive output (middle) and negative output (bottom) signal waveforms for the TPA-3113D2 Class-D audio amplifier IC.

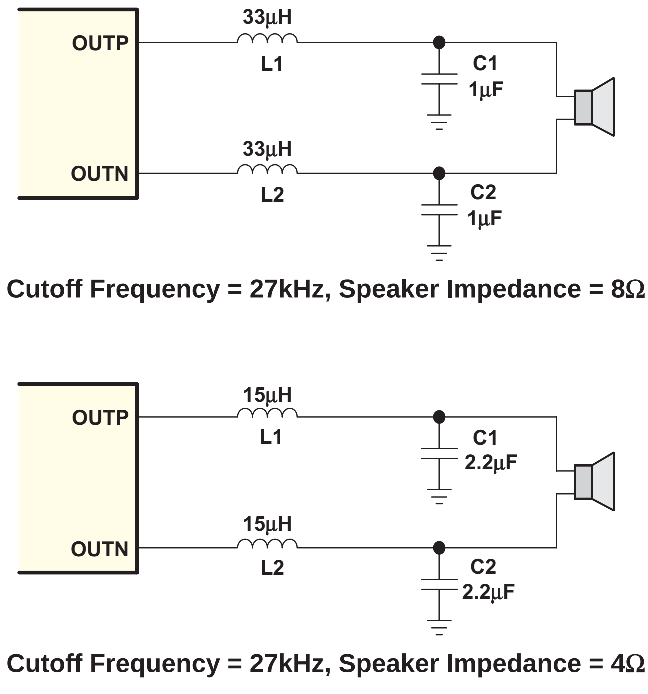


Fig.3: an external LC filter can be added if long speaker leads are to be used, with the filter component values selected according to the speaker impedance. These two diagrams show the recommended values for 8Ω and 4Ω loads.

an external LC output filter. This could be wired externally to the board, ie, between the output terminals and speakers. Note that you would need to keep the components relatively close and run some connections to a PCB ground point.

One drawback of this approach is that filter component values must be selected based on the speaker impedance. Also, the inductors must handle the peak load current (up to 4A in some cases) without saturating. The recommended filters for 8Ω and 4Ω loads are shown in Fig.3. Note that an LC filter may also give improved efficiency.

Speaker impedance

For supply voltages up to 15V, the unit can drive speakers with nominal impedances from 4-8Ω. Above 15V, however, it isn't recommended to drive 4Ω speakers. Plenty of power for 4Ω loads is already available at supply voltages below 15V anyway.

To drive 4Ω speakers from a supply above 15V, it's necessary to run the *Mini-D* in mono mode; more on that later. To drive two speakers in this mode, you will need to build two boards but in exchange for that, you get more power and higher efficiency.

Circuit description

The full circuit is shown in Fig.4. All the real work is done by IC1. The left and right channel input signals are applied to RCA connectors CON1 and CON2. Alternatively, pin headers may be fitted in their place for connection to chassis sockets or another board. From this point on, we shall refer to the operation of one channel only.

The signal first passes through a low-pass RF-rejecting filter, comprising a 100Ω series resistor and 1nF ceramic capacitor. Both the signal and ground pins are then AC-coupled to the volume control potentiometer (VR1) via 4.7µF ceramic capacitors. The signal ground is also connected to power supply ground via a 4.7Ω resistor, taking advantage of the differential inputs provided by the IC.

This 4.7Ω resistor reduces the chance of hum being injected into the signal due to the common input grounds.

The volume control potentiometer is either a dual-gang log pot (VR1) or two horizontal trimpots (VR2 and VR3), the latter used for a pre-set volume level. If you don't need volume control at all, simply link out VR2 and VR3. Regardless, the wiper of each pot goes to the non-inverting input for each channel (pins 3 and 12) while the bottom (AC-grounded) end goes to the inverting inputs (pins 4 and 11).

The TPA3113D2 can handle a strictly limited voltage range at each input pin of –0.3V to 6.3V, so we have added protection components to limit these voltages when the power is off or in case a high-level signal is applied (which is common when plugging and unplugging RCA leads).

These parts consist of 5.6V zener diodes (ZD1-ZD4) and parallel Schottky diodes (D1 and D2) between each input and ground. The zener diodes take care of clamping positive signal swings, while the Schottky diodes clamp negative excursions more effectively.

The outputs of IC1 pass through the recommended output filter, consisting of four large ferrite beads (FB1-FB4; HI1812V101R-10) and four 1nF C0G ceramic capacitors. C0G capacitors have a very low temperature coefficient (±30ppm) but also low ESR (equivalent series resistance) and ESL (equivalent series inductance); just what we need to suppress sharp voltage spikes.

We have also added snubbers, consisting of 330pF C0G ceramic capacitors

Specifications

Supply voltage: 8-25V DC

Quiescent current: typically <40mA active, <2mA shutdown

Speaker impedance: 6-8Ω; 4-8Ω mono mode or stereo with up to 16V supply

Continuous output power: 2 x 5W or 1 x 10W (12V, 8Ω)

Peak output power: 2 x >15W or 1 x >30W (thermally limited)

THD+N: typically <0.1%; see Figs.6 and 7

Signal-to-noise ratio: 100dB

Frequency response: 20Hz-20kHz ±1dB; see Fig.9

Efficiency: up to 82% (stereo), 91% (mono)

Gain: 20dB, 26dB, 32dB or 36dB

Under-voltage lockout: ~7.5V

Output offset voltage: typically within ±1.5mV

Power supply rejection ratio: typically -70dB

Switching frequency: ~310kHz

in series with 10Ω resistors, from each output to ground. They are actually wired to the boost supply pins, but these are AC-coupled to the outputs via much larger 220nF capacitors, so the effect is the same. These reduce radiated EMI further by limiting the output voltage slew rates.

We have used a 1:1 voltage divider between GVDD (pin 9; ~7V) and ground, with a 4.7µF filter capacitor, to set PLIMIT (pin 10) at 3.5V. This limits the output amplitude to about ±11V (22V peak-to-peak). Thus it will only limit the output power with a DC supply over 20V.

If you are trying to get the maximum possible power from the chip at 24V, you could reduce the upper divider resistor to 47kΩ, but in most cases it won't make much difference; the 'soft clipping' provided by this limiter may have some benefits in reduced treble artefacts if you are going to drive the amplifier that hard anyway.

Other features

100kΩ pull-ups on GAIN0 and GAIN1 allow links LK4 and LK5 to define these input states. A table in the circuit diagram shows the possible settings. With a gain of 20dB (10x), input sensitivity is 425mV RMS for a 12V supply and 850mV RMS for a 24V supply. With the gain set to 36dB (63x), input sensitivity is 67mV RMS for a 12V supply and 135mV RMS for a 24V supply.

The unit can handle signals up to at least 3V RMS. For line-level signal sources such as CD players, 20dB of gain should be plenty, so most constructors should stick with that.

The FAULT output (pin 2) is connected to pin 1 on CON6, which can go to a microcontroller pin (but with some provisos, see below). It goes low if the IC detects that an output is short-circuited or there is a DC offset fault. The FAULT signal also switches P-channel MOSFET Q2 via a resistive divider (which ensures that Q2's gate is not over-driven). If there is a fault, Q2 switches LED2 (red) on. This can either be an SMD LED mounted on the board or an external LED wired up via pin header CON8.

The shut-down input (pin 1) is also connected to CON6 (at pin 2) and is pulled up by a 100kΩ resistor so that the amplifier will power up automatically. If pulled to ground, the amplifier shuts down and only draws about 250µA. However, that doesn't include the current for LED1 and the various pull-ups, which increase total shutdown current to around 2mA.

If a shorting block is placed on LK6 and an output short circuit is detected, once the short has cleared, the amplifier will automatically resume operation. Otherwise, short-circuit faults are 'latched' and the unit remains off (with LED2 lit) until the power is turned off and back on again.

Over-temperature faults are automatically cleared and LED2 will not light if IC1 overheats; rather, output will simply cease and then resume once it has cooled.

Power LED1 (green) can either be an on-board SMD LED or it can be mounted off-board via pin header CON7. Supply current for LED1 and LED2 is around 1-2mA, so high-brightness types should be used.

Power supply

The 8-25V DC supply (from a battery, plugpack or power 'brick') comes in via terminal block CON3, with MOSFET Q1 providing reverse polarity protection. If the supply polarity is correct, Q1's gate is pulled positive via the 100kΩ resistor. This switches Q1 on, so current from the circuit can flow back to the supply ground.

However, if the supply polarity is wrong, Q1's gate will be pulled negative relative to its source and Q1 will remain off, so no ground current can flow and the circuit is protected. Q1's drain-source voltage is rated at 25V, so as long as the DC supply is within the specified range, this will be sufficient to block the supply voltage.

Zener diode ZD5 limits Q1's gate voltage to a safe level when the supply voltage is above 20V.

There's little else to the power supply other than the bypass capacitors, which consist of one 100µF electrolytic, one 220nF X7R ceramic and one 1nF C0G ceramic for each pair of power V_{CC} pins, ie, PVccL (pins 27 and 28) and PVccR (pins 15 and 16). The analogue supply, AVcc, is at pin 7 of IC1 and has a 10Ω/4.7µF RC low-pass filter to remove switching noise.

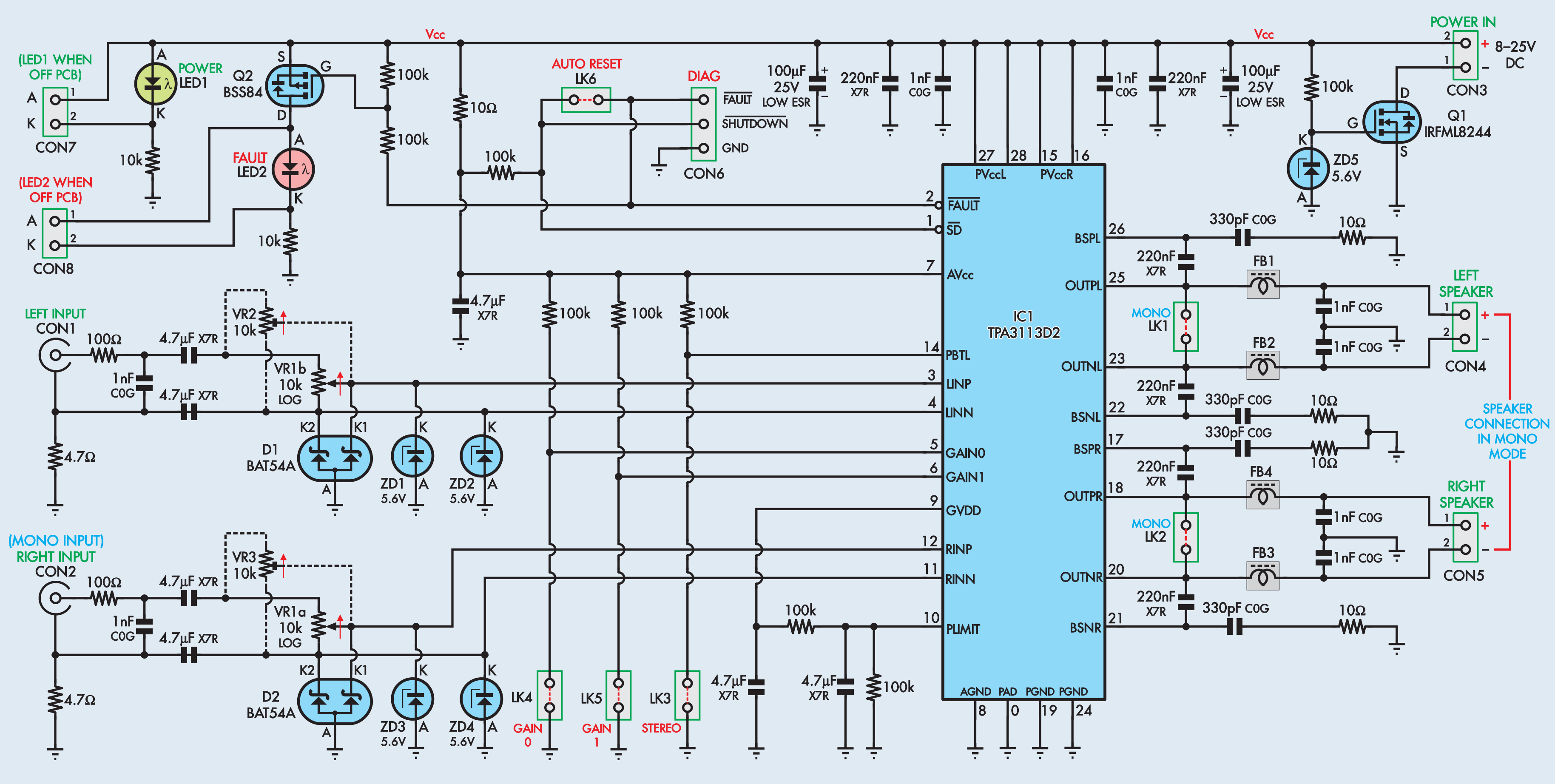
IC1's internal MOSFET gate supply regulator has a 4.7µF output filter capacitor at pin 9 (GVDD).

Mono (parallel) mode

To operate in mono mode, LK1 and LK2 are fitted and LK3 is left out. The speaker is then connected between CON4 and CON5, as shown on the circuit diagram. LK1, LK2 and LK3 are 0Ω surface-mount resistors.

In this case, you can also omit FB2, FB3 and the two associated 1nF capacitors. Plus you can omit CON1 and its associated components because the mono signal is fed into the right input (CON2).

Note that you will only get more power in mono mode (also known as PBTL or 'Parallel Bridge-Tied Load' mode) with a low-impedance speaker, eg, 4Ω. This is because with higher speaker impedances, you will run into clipping before the maximum output current becomes the limiting factor. With a 4Ω speaker at 15V in mono mode, output power is



LK4	LK5	GAIN
IN	IN	20dB
IN	OUT	26dB
OUT	IN	32dB
OUT	OUT	36dB

STEREO/MONO MODE LINKING			
MODE	LK1	LK2	LK3
STEREO	OUT	OUT	IN
MONO	IN	IN	OUT

* EITHER DUAL LOG POT VR1 OR TRIMPOTS VR2 & VR3 ARE FITTED; THE LATTER FOR PRESET VOLUME

NOTE: FB1–FB4 ARE HI1812V101R-10 SMD FERRITE 'BEADS' (1812 SIZE, 8A RATED)

'MINI-D' 10W×2 CLASS D AMPLIFIER

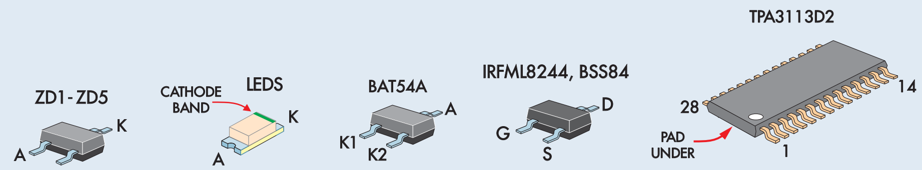


Fig.4: the complete circuit diagram of the *Mini-D 10W x 2 Class-D Amplifier*. The audio input signals are fed in via CON1 and CON2, filtered and fed to the LINP (pin 3) and RINP (pin 12) inputs of IC1 via dual-gang volume control VR1. The amplified outputs appear at pins 25 and 23 (left channel) and pins 18 and 20 (right channel) of IC1, and these drive the left and right speakers via ferrite bead/capacitor filter networks. Links LK1-LK3 select either mono or stereo operation (see table), while links LK4 and LK5 set the gain. MOSFET Q1 provides reverse polarity protection, while MOSFET Q2 drives the fault indicator LED (LED2).

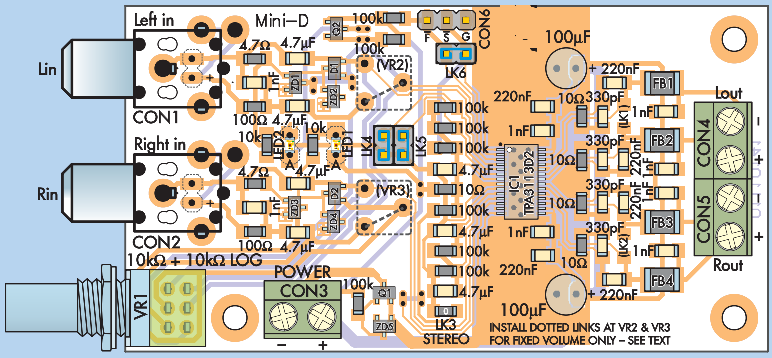
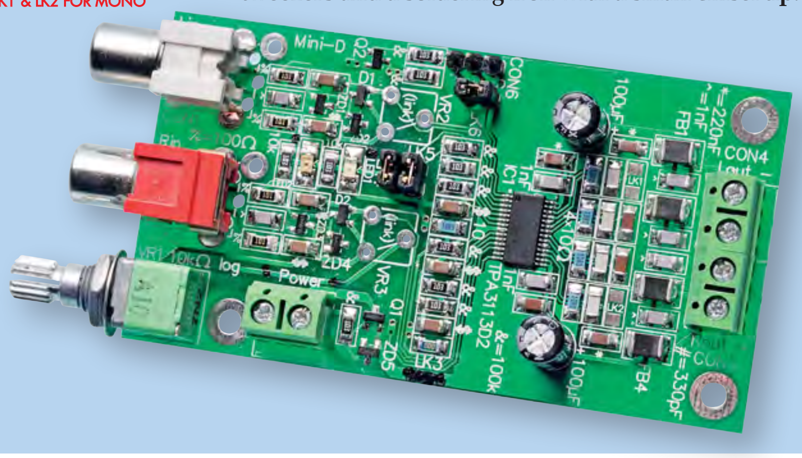


Fig.5: follow this parts layout diagram to build the *Mini-D* amplifier. You can either install potentiometer VR1 or trimpots VR2 and VR3 for volume control (see text). Alternatively, leave all these parts out if no volume control is required, and link out VR2 and VR3 as indicated.



Below: the completed PCB assembly. Don't be intimidated by the SMD parts; they're quite easy to install if you follow the instructions in the text, but you do need a good magnifying glass (or magnifing lamp), tweezers and a soldering iron with a small chisel tip.

up to 30W, which is pretty good! Even if you don't need the extra power, it's preferable to use the module in mono mode as it improves efficiency.

24V battery operation

Since the maximum recommended operating supply voltage for IC1 is 26V and there are a number of 25V-rated components in the circuit, we don't recommend running directly from a 24V battery. In theory, if you increased the voltage ratings of the 25V capacitors and MOSFET Q1, you might get away with it, as the absolute maximum specified for IC1 is 30V. But it's outside the recommended operating voltage range so we don't suggest doing that.

A better option is to use a 24V low-dropout pre-regulator, eg, by placing a 12V zener diode in series with the ground pin of an LM2940CT-12 regulator to 'jack it up' to 24V. You will need appropriate input and output filter capacitors. The LM2940 is only rated at 1A, but is unlikely to run into current limiting during normal operation. It may need a small heatsink though, as it could dissipate up to 5W.

PCB layout

Being a switching amplifier, instantaneous currents can be high and the voltage rise/fall times are very short, so the the design of the PCB has been quite rigorous. We also wanted to keep switching noise away from the analogue circuitry. Bypass capacitors need to be near IC pins and the output filter must be kept tight for maximum EMI suppression. There are also thermal considerations, given that the amplifier IC uses the board as a heatsink.

We've placed ground planes on both the top and bottom of the board immediately under IC1 and fanned them out to the full width of the board. There are 15 vias placed directly under the IC, on and around its thermal pad, both to reduce ground impedance for better performance and to help conduct heat from the IC to the bottom side of the board, where it can be effectively radiated away.

The 1nF and 220nF bypass capacitors are immediately adjacent to the IC, with the 1nF C0G types the closest, as they have the best high-frequency performance. The placement of the 100µF electrolytics is less critical. Note that there is provision to use 22µF 25V SMD multi-layer ceramic capacitors (1812 size) instead, but the cheaper electros do the job well.

The IC's pin layout is well-optimised, with the main power supply and all output-related pins on one side, which we have oriented towards the right side of the PCB. Thus the filter components are placed immediately between the IC and CON4/CON5 at right. The analogue ground pin (pin 8) is on the left side of the IC and this is the only point at which the power ground meets the signal (analogue) ground.

Construction

Fig.5 shows the assembly details. Apart from some of the components being relatively close together, the only tricky thing about building this board is soldering IC1 (a magnifying lamp will come in handy here).

We used hot-air reflow as this (or oven reflow) is best for ICs with thermal pads (like the TPA3113D2). The equipment is surprisingly cheap; we paid around £35 for an Atten 858D+ hot-air soldering station, while hot-air reflow wands can be had for as little as £15. But you can do it with a regular soldering iron too.

For hot-air, the trick is to use a very thin layer of fresh solder paste (kept in the fridge!). Spread this sparingly on the pads, drop the IC on top, heat it (gently at first) until all the pins reflow and then for a few seconds longer and 'Bob's your uncle'.

If all you have is a regular iron, apply some no-clean flux paste to the thermal pad on the board and also the pad on the bottom of the IC. Then melt a small amount of solder to both; just enough to tin them. Start with the PCB pad, so you can get an idea of the correct amount. If you add too much, add a bit more flux and then remove the excess with some solder wick (harder to do with the IC!).

Having tinned both, place some fresh flux paste on all the IC pads on the PCB, including the thermal pad, then pop the IC down in place,

checking its orientation. Next, move it slightly out of the way, tin one small corner pad and then slide the IC into place while heating that pad.

Now check that the IC lines up with all its pads. If it's misaligned, reheat and gently nudge it into place. Try to avoid getting solder on any other pads

Use a magnifying glass (or magnifying lamp) to check carefully that all the pins are sitting properly over their pads, then tack down the diagonally opposite pin. Re-check the alignment, then solder the rest of the pins, making sure not to disturb either of those first two solder joints.

Having soldered the pins, it will then be necessary to flip the board over and apply enough solder to the pad on the bottom to transfer heat through the vias. Heat this solder until the flux between the IC and board vapourises, indicating that the thermal pad has reflowed. This will take a good few seconds but don't overdo it as you could cook the chip.

Regardless of which method you used to solder the IC, check carefully for bridged pins (again, use a magnifying glass) and clean any that look dodgy with some flux paste and a clean piece of solder wick. The bridges should clear easily; press the wick down onto the board – but not over the IC pins – as they are small and easily damaged.

As a final measure, it's a good idea to clean the flux residue off the board using a specialised flux cleaner (or at a pinch, an alcohol or acetone) and then carefully check all the soldering, again with a magnifying glass. Check that all the bridges are gone and that the solder has flowed cleanly onto all the pins and pads.

Remaining parts

There are nine SOT-23 package transistors, diodes and zener diodes to solder. These are quite easy as the pins are well spaced – but don't get the various device types mixed up. Start with Q1 and Q2, then solder D1 and D2, and finally the five identical zener diodes. The easiest method is to put a bit of solder on the central pad and slide the device into place while heating that pad. Then solder the other two pads (a dab of flux paste makes it easier) and refresh the first.

Now move onto the SMD passives, starting with the resistors and then the capacitors and ferrite beads. Use a similar method as for the SOT-23s.

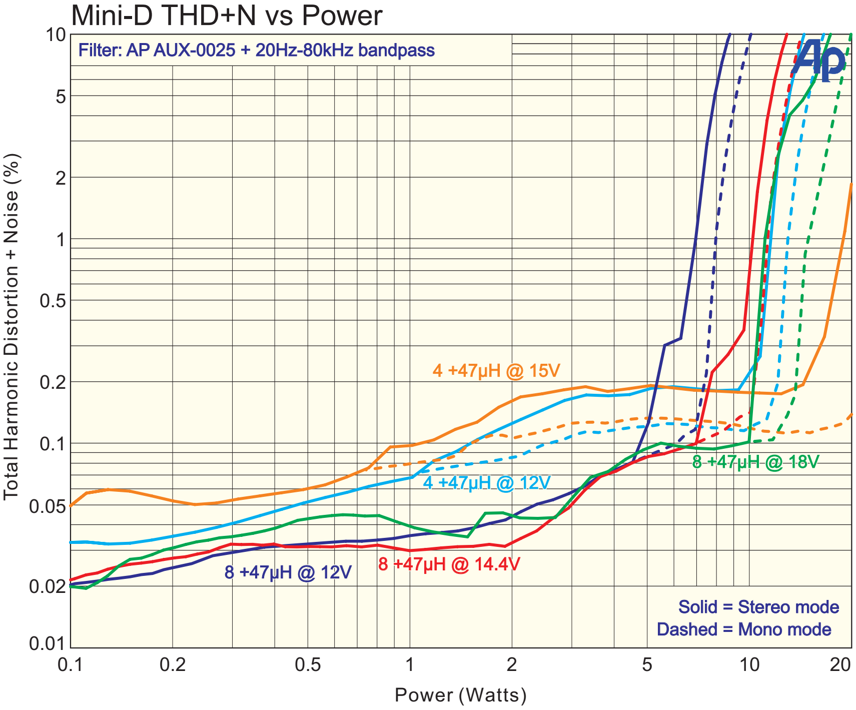


Fig.6: distortion versus power for a range of load impedances and supply voltages. Performance is generally better for 8Ω loads, but power delivery is higher into 4Ω. Note the test load series inductance, to simulate loudspeakers.

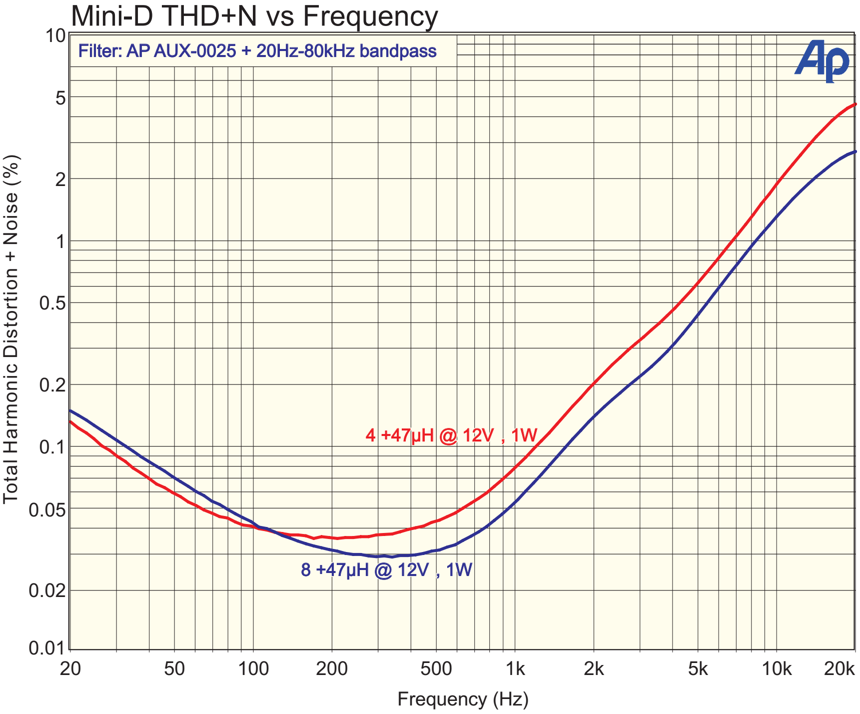


Fig.7: distortion versus frequency at 1W. As usual, the distortion rises with frequency, but it also rises at the low end due to coupling-capacitor-induced distortion. It's below 0.1% between 40Hz and about 1.5kHz.

The main thing to check for with these parts is that the solder has flowed onto the pad and not just the end of the component. As before, adding flux smoothes solder flow.

Note that the resistors will have printed values on them, but the capacitors and ferrite beads will not, so check the packaging before fitting them. Remember to fit either 0Ω resistor LK3 only (stereo mode) or LK1 and LK2 only (mono mode).

If using the SMD LEDs, they can go in next, but first you will have to check

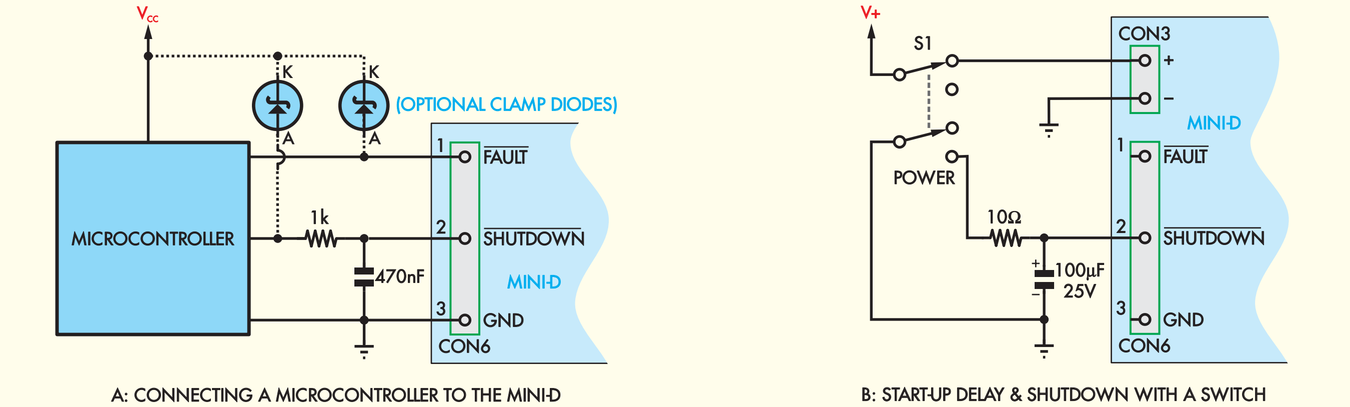


Fig.8(a): the shutdown pin (pin 2) of CON6 can be pulled low under no-signal conditions (eg, using a microcontroller) to reduce power consumption. The RC filter shown provides slew-rate limiting, while external clamp diodes may also be required with some micros (see text). Fig.8(b) at right shows how to add a capacitor (eg, 100µF) to give a switch-on delay, while a DPDT power switch (S1) can be used to eliminate switch-off clicks or pops.

their polarity. Unfortunately, markings are inconsistent so use a DMM in diode test mode and try connecting the probes both ways around. When the LED lights, the red probe is to the anode and this goes towards the bottom of the PCB (marked with 'A'). We used a green LED for LED1 and a red LED for LED2.

Through-hole components

That's it for the SMDs, so once you're confident that they've all been soldered correctly, there are just a few through-hole parts left. If you aren't using an on-board volume control, solder wire links in place of VR2 and VR3 where shown. Also, if using off-board LEDs, fit 2-way pin headers CON7 and CON8 in place of the LEDs.

Next, move on to links LK4-LK6, CON6 and the inputs (if you aren't fitting RCA sockets). That done, dovetail two screw terminal blocks together and solder them in place for CON4 and CON5 (wire entry holes facing outwards). CON3 can then go in.

If you are using onboard volume control pot VR1, fit it now (or trimpots VR2 and VR3). RCA sockets CON1 and CON2 can then go in, followed by the electrolytic capacitors (take care with their orientation).

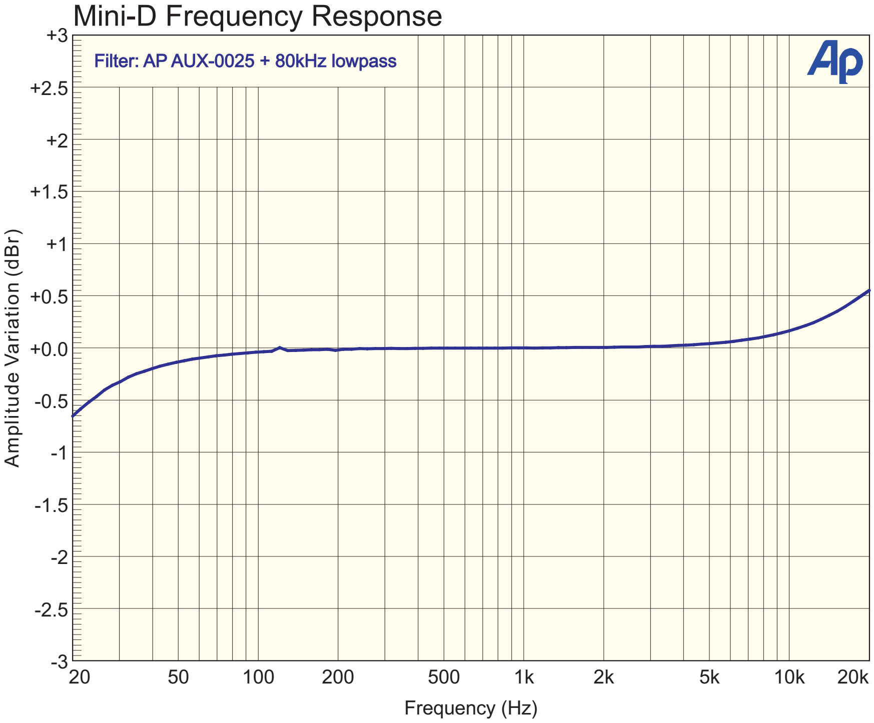


Fig.9: the amplifier's response is effectively flat in the audible frequency range. There is a low-frequency roll-off due to the high-pass filter formed by the input coupling capacitors and volume pot, while the rise at the high end can be attributed at least partially to the inductance of our test load.

Set-up and testing

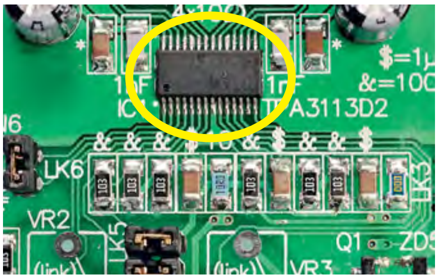
Initially, fit LK4 and LK5 (note that they go in vertically) and LK6. Turn the volume pot(s) to minimum, then apply DC power to CON3 (say, 12V) and measure the current. It should be just under 40mA (but possibly as high as 55mA) and LED1 should be on while LED2 should be off. If anything is wrong, switch off immediately and check for faults. Also, double-check that you have connected the supply wires with the correct polarity.

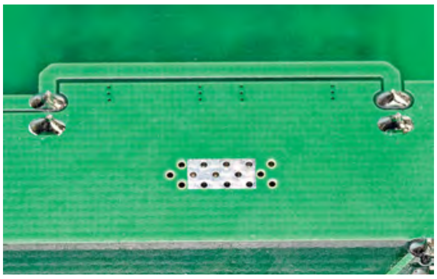
Assuming that all is OK, switch off and connect a signal source such as a CD player, MP3 player, oscillator or mobile phone. Connect the speaker(s), then switch back on and slowly turn the volume up. It's now just a matter of making sure it sounds right. If you get to maximum volume and it's still too quiet, switch off and increase the gain by changing LK4 and/or LK5, but do remember to turn the volume down before re-applying power.

Shutdown control

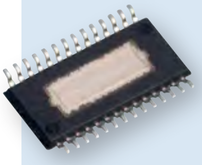
To reduce power consumption when power is applied but no signal is present, you can pull the shutdown input (pin 2 of CON6, pin 1 of IC1) low to enter a power-saving state. However, there are a couple of provisos.

First, the data sheet specifies that this pin should be slew-rate limited to 10V/ms unless the source impedance is at least 100kΩ, but it doesn't say why. Confusingly, they also show





The TPA3113D2 IC (circled) should be installed first, following the procedure described in the text. The photo above right shows the corresponding heatsink area for this IC on the back of the PCB. It's connected to a thermal pad on the top of the board by 15 vias.



This larger-than-life-size view shows the heatsink pad on the underside of the TPA3113D2 Class-D audio amplifier IC.

Another view of the completed PCB assembly. Links LK1-LK3 have been configured for stereo operation; ie, LK1 (0Ω) in, LK2 and LK3 out.

sample circuits where a 'control system' (eg, a microcontroller) drives the shutdown pin via just a 1kΩ series resistor, which is unlikely to limit the slew rate to their specification.

We would be tempted to try that, but not knowing the reason for the limitation, a safer approach would be to add an RC filter, as shown in Fig.8(a). The same comments apply if you're going to use a switch, relay, transistor or something else to pull down the shutdown pin.

If connecting a micro in this manner, note that the on-board pull-up resistor could pull its control pin above the micro's supply voltage. Normally, the microcontroller pin will have a clamp diode to its positive supply rail to limit the voltage on that pin to a safe level. However, some micros lack a positive clamp diode (eg, 5V-tolerant pins on a 3.3V micro) and in that case, you will need to add an external clamp diode (or a low-voltage zener to ground) to protect the micro – see Fig.8(a). The situation is the same if connecting the FAULT signal to a micro.

Powering up and down

We didn't hear any clicks or pops or run into other issues when powering the *Mini-D* up or down normally, but there are a couple of issues noted in the data sheet which constructors should be aware of.

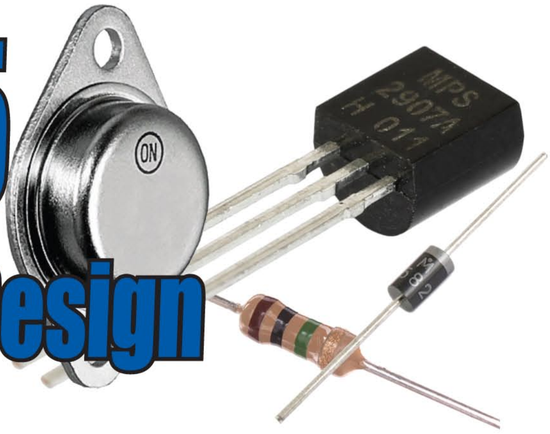
If the signal source is powered up at the same time as the *Mini-D* and there are large initial transients on those signals, that could trigger the DC offset protection in the *Mini-D,* then once that's activated, its outputs will remain disabled until the power is switched off and on again. So in that case you need to hold shutdown low until the audio signals stabilise. This can be achieved with a capacitor between the shutdown pin and ground. A 22µF capacitor will give a switch-on delay of around 100-200ms, a 100µF 500-1000ms and so on. Or if a micro is connected to shutdown, it can do the same job.

The data sheet also states that pulling shutdown low before power is removed will minimise clicks or pops. While not strictly necessary, this can be achieved using a DPDT power switch; see Fig.8(b). This will bring shutdown low almost immediately while the supply capacitors take some time to discharge.

Teach-In 2015

Discrete Linear Circuit Design

Part 8: Power amplifiers



by Mike and Richard Tooley

Welcome to *Teach-In 2015*. This series is aimed at anyone wishing to develop a detailed understanding of linear discrete semiconductor devices and how they are used in a diverse range of circuits. We hope you will join us on this exciting voyage of discovery! Each part of our *Teach-In 2015* series is devoted to a different aspect of discrete linear circuit design such as modelling and simulation, measurement and testing, and noise and distortion. In last month's instalment, *Knowledge Base* introduced you to three more useful circuit building blocks in the form of the current mirror, the differential amplifier and the V_{BE} multiplier. *Get Real* described the construction and use of a simple VU-meter, while *Discover* was devoted to heat and heat dissipation with the necessary theory to help you select the right heat sink for your own designs.

Introduction

In this month's *Teach-In 2015*, *Knowledge Base* introduces negative feedback and explains how this provides a useful way of making an amplifier both stable and predictable. *Discover* is devoted to output stages and introduces a number of useful circuits including *Darlington* output stages, compound feedback *Sziklai* pairs, and the quasi-complementary configurations that are commonly used in them.

Knowledge base: Negative feedback

Many practical amplifiers use negative feedback in order to precisely control the gain, reduce distortion and improve bandwidth. The gain can be reduced to a manageable value by feeding back a small proportion of the output. The amount of feedback determines the *overall* (or *closed-loop*) gain. Because this form of feedback has the effect of reducing the overall gain of the circuit, this form of feedback is known as *negative feedback*. An alternative form of feedback, where the output is fed back in such a way as to reinforce the input (rather than to subtract from it) is known as *positive feedback*. Positive feedback has the effect of reducing the stability of an amplifier and, when the overall loop gain exceeds unity oscillation can result (an undesirable consequence in the case of an amplifier).

Fig.8.1 shows the block diagram of an amplifier stage having internal voltage gain A, to which negative feedback has been applied. Thus, when just considering the *amplifier stage alone* we can say that:

$$A = \frac{V_{out}}{V'_{in}} \qquad (1)$$

Note that the voltage that appears at the input of the amplifier stage is not the same as the input voltage applied to the overall circuit when negative feedback has been applied – the *overall* voltage gain will be given by a different expression:

$$G = \frac{V_{out}}{V_{in}} \qquad (2)$$

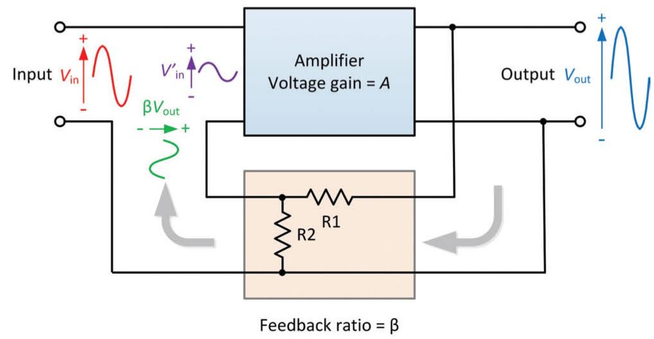


Fig.8.1 Negative feedback principle

The proportion of the output voltage fed back to the input is given by β, such that the voltage appearing at the output of the feedback network (referred back to the input) is:

$$V_{FB} = \beta V_{out}$$

The input voltage to the amplifier stage, V'_{in}, is the *difference* between the overall input voltage, V_{in}, and the voltage fed back, V_{FB} (note that the amplifier's input voltage has been effectively *reduced* by applying negative feedback), thus:

$$V'_{in} = V_{in} - \beta V_{out}$$

Rearranging this expression gives:

$$V_{in} = V'_{in} + \beta V_{out} \qquad (3)$$

Combining equations (2) and (3) gives:

$$G = \frac{V_{out}}{V'_{in} + \beta V_{out}} \qquad (4)$$

Combining equations (4) and (1) gives:

$$G = \frac{V_{out}}{\left(\frac{V_{out}}{A}\right) + \beta V_{out}} = \frac{A}{1 + \beta A}$$

The foregoing relationship tells us that the overall gain with negative feedback applied will always be less than the gain without feedback (in other words, G is always less than A). Furthermore, if A is very large then the overall gain with negative feedback applied will be approximately equal to 1/β. Note, also, that the loop gain of an amplifier with feedback applied is defined as the product of β and A.

To put this into context, let's assume that we are dealing with an amplifier that has a gain (without any feedback applied) of 25 and that 6% of its output is fed back to the input as negative feedback, as shown in Fig.8.2. We can determine the resulting closed-loop voltage gain using:

$$G = \frac{A}{1 + \beta A}$$

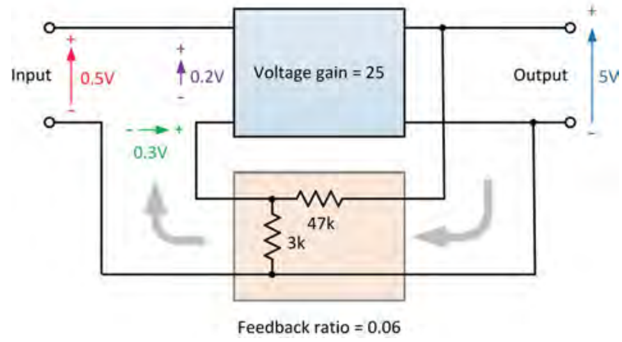


Fig.8.2 Negative feedback example

Table 8.1 Effect of feedback method on input/output resistance

Feedback method	Fig.8.3	Input resistance	Output resistance
Voltage derived, parallel applied	(a)	Decreased	Decreased
Voltage derived, series applied	(b)	Increased	Decreased
Current derived, parallel applied	(c)	Decreased	Increased
Current derived, series applied	(d)	Increased	Increased

In this case $A = 25$ and $\beta = 0.06$ (ie, 6%) resulting from the potential divider formed by the two resistors present in the feedback path. Thus, the resulting amplifier stage with negative feedback applied will have a voltage gain of:

$$G = \frac{25}{1+(0.06\times 25)} = \frac{25}{1+1.5} = 10$$

Advantages of negative feedback

Negative feedback confers a number of advantages – importantly, it can help us define and stabilise the gain of an amplifier when it would otherwise be susceptible to variations in component values and semiconductor parameters. Negative feedback also helps reduce distortion and increase bandwidth.

Since transistor current gain can often vary over quite a wide range, gain stabilisation is important. It's worth putting this into context with an example showing how the closed-loop gain in the previous example will change if the gain of the amplifier stage varies. Let's suppose that, due to variations in transistor current gain, the internal gain of the amplifier used earlier can range from 20 to 30 (ie, 25 ± 5, or ±20%). With 6% negative feedback applied to the amplifier the voltage gain will only range from about 9.09 to 10.71 (i.e. less than a ±10% variation).

Applying negative feedback

Negative feedback can be applied in several different ways. In a conventional voltage amplifier like those that we've already met in this series, the feedback voltage can be derived from either the voltage or the current present at the output. This voltage can then be fed back to the input so that it either appears in parallel or in series with the applied input voltage.

Thus, we have four different ways in which negative feedback can be applied to voltage amplifiers:

(a) Feedback derived from the output voltage and applied in parallel with the input voltage, Fig.8.3a

(b) Feedback derived from the output voltage and applied in series with the input voltage, Fig.8.3b

(c) Feedback derived from the output current and applied in parallel with the input voltage, Fig.8.3c

(d) Feedback derived from the output current and applied in series with the input voltage, Fig.8.3d

In this general introduction to negative feedback (and in the practical circuits that follow) we will confine our description to methods (a) and (b) shown respectively in Fig.8.3a and Fig.8.3b.

Input and output impedance

As well as reducing distortion, stabilising gain, and increasing bandwidth, negative feedback has an impact on the input and output impedance of an amplifier. For example, with voltage-derived feedback, cases (a) and (b), the output resistance will be decreased, but with current-derived feedback, cases (c) and (d), the output resistance will be increased. Conversely, with parallel-applied feedback, cases (a) and (c), the input resistance will be decreased but with series-applied feedback, cases (b) and (d), the input resistance will be increased. We've summarised all of this in Table 8.1.

Practical negative feedback arrangements

To put all of this into a practical context, Fig.8.4b and Fig.8.4c show how negative feedback can be very easily implemented in a single-stage transistor amplifier. For comparison, a common type of single-transistor amplifier with DC bias

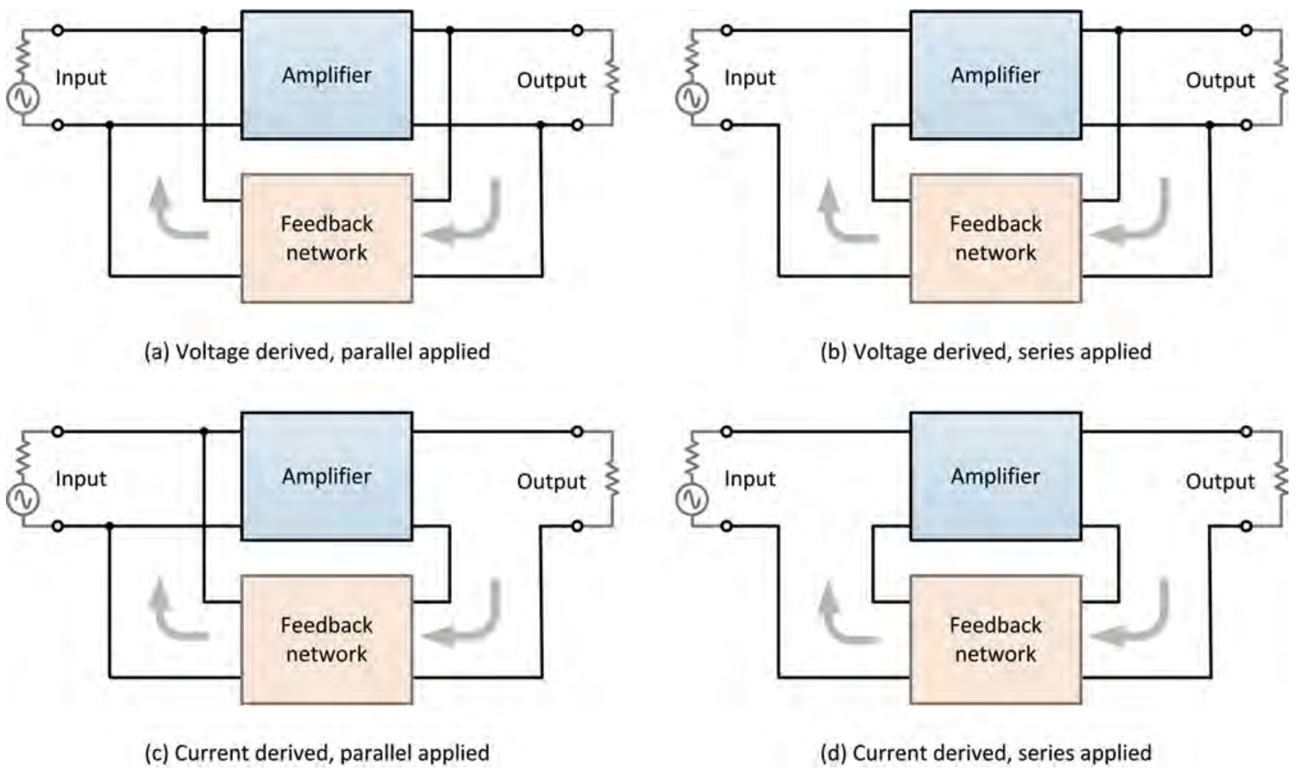
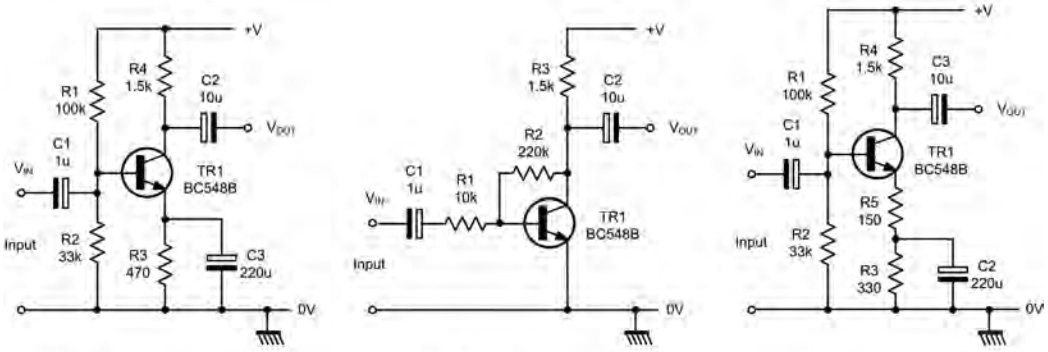


Fig.8.3 Four different ways of applying negative feedback



(a) Conventional stabilised amplifier (b) Shunt voltage feeback (c) Series current feeback

Fig.8.4 Simple ways of adding negative feedback to a single-stage transistor amplifier

stabilisation but *without* AC feedback is shown in Fig.8.4a.

In Fig.8.4b the common-emitter amplifier has feedback applied from its output (the collector) to its input (the base). The feedback resistor, R2, supplies the DC base bias current for the transistor while, at the same time, it also provides AC signal feedback from collector to base. The feedback is negative because the signals at the collector and base are 180° out of phase. Since the output signal voltage is fed back so that it appears in parallel with the input, this type of negative feedback is often referred to as *shunt voltage feedback* (see method (a) described previously).

The amount of negative feedback in Fig.8.4b is largely determined by the ratio of R2 to R1 but, since the source impedance (not shown in Fig.8.4) also appears in series with R1 this also has an effect on the closed-loop voltage gain. With the values shown, and assuming that the signal source impedance is relatively low, the closed-loop voltage gain will be between about 10 and 15. You can check this out easily using Tina Design Suite, as shown in Fig.8.5.

In Fig.8.4c the common-emitter amplifier has negative feedback applied by leaving the emitter circuit un-bypassed. This results in a signal voltage appearing at the emitter (as well as at the collector). The feedback voltage is derived from the emitter current of the transistor and applied in series with the input signal voltage and so this form of feedback is often referred to as *series current feedback* (see method (b) described previously).

The amount of negative feedback in Fig.8.4c is determined by the ratio of the value of collector load resistance, R4, to the un-bypassed emitter resistance, R5. With the values shown, and once again assuming that the signal source impedance is relatively low, the closed-loop voltage gain will be between about 7.5 and 10. Once again, you can easily verify this using Tina Design Suite, as shown in Fig.8.6.

As mentioned earlier, apart from affecting the voltage gain of an amplifier, there are other important consequences of introducing negative feedback to these simple amplifier stages. For example, in the case of shunt voltage feedback in Fig.8.4b the overall input impedance of the stage will become reduced, while, in the case of series-current feedback in Fig.8.4c, the overall input impedance of the stage will be increased.

Controlling the gain

The gain of the shunt voltage feedback circuit shown in Fig.8.4b can be very easily changed by varying the ratio of R2 to R1. In practice, R2 will be selected in order to provide the correct value of base bias for TR1 and practical values for this component usually lie in the range 100kΩ to 1MΩ. With the value of R2 shown, a collector current of about 3mA will be produced (corresponding to a base current of a little less than 20µA). Altering the value of R1 will have no effect on the bias applied to TR1, but it will affect the feedback ratio and in turn the voltage gain. If the value of R1 is increased the voltage gain will fall, and vice versa.

In the circuit shown in Fig.8.4c, the combined emitter resistance is made up from the series resistance of two components, R3 and R5. Note that R3 is bypassed by means of C2, as in Fig.8.4a, but an additional resistance, R5, has been added in order to introduce series current feedback. DC bias for the emitter is established by means of the series combination, R3 + R5, and so the stabilised bias conditions are very similar to those shown in Fig.8.4a.

The voltage gain of the circuit shown in Fig.8.4c is determined by the ratio of R4 to R5. With R4 fixed, reducing the value of

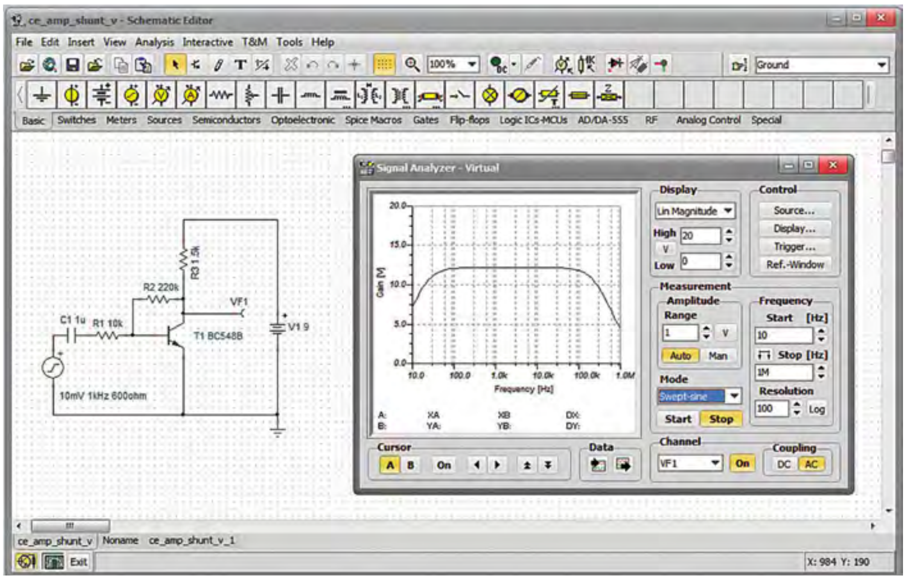


Fig.8.5 Testing shunt voltage feedback in a single-stage transistor amplifier

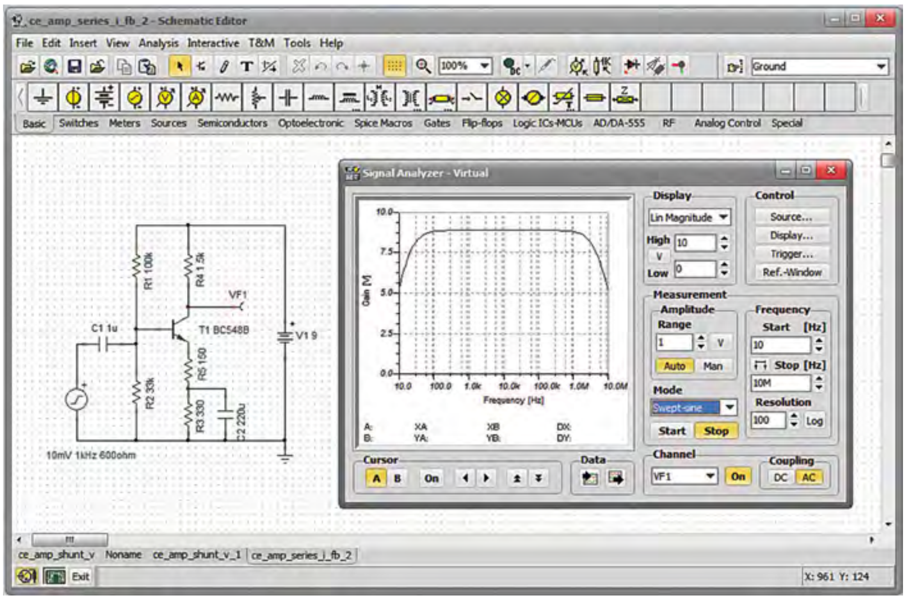


Fig.8.6 Testing series current feedback in a single-stage transistor amplifier

R5 (and correspondingly increasing R3 so that the total resistance, R3 + R5, present in the emitter circuit remains constant) will have the effect of increasing the voltage gain of the stage, and vice versa.

With all this in mind, it can be a useful exercise to modify the test circuit arrangements shown in Fig.8.5 and Fig.8.6 to produce a specific value of voltage gain. We suggest that you try to obtain a voltage gain of 20 from each of the two circuits. When you do this, don't forget to use Tina Design Suite's DC analysis tools to check that each stage remains correctly biased!

Feedback over several stages

So far, we have only considered the application of negative feedback over a single stage, but it is common practice to apply feedback over several stages as we will see next month in Part 9. In multi-stage amplifiers it is usual for the open-loop gain (the gain before applying feedback) to be quite large, but attention also needs to be given to the overall phase shift. This needs to ensure that not only is the feedback applied in the correct sense (ie, that it is negative rather than positive) but also that the amplifier will remain unconditionally stable over its entire frequency range.

It is important to note that, while the phase shift produced by an amplifier will remain substantially constant over the mid-band range, due to the presence of multiple reactive components the phase shift will often be subject to considerable variations at extremes of frequency. This can, in turn, result in continuous oscillation at frequencies that are well outside the normal pass-band of the amplifier as the feedback becomes positive rather than negative. With this in mind, designers often incorporate networks of passive components in order to compensate for undesirable phase shifts. We will be delving a little deeper into this complex problem next month.

Discover: Increasing output power

Earlier in the series we explained how the properties of an emitter-follower stage make it ideal for use as the final stage of a medium or high power amplifier (see *Teach-In 2015* Part 3, April 2015). Our *Get Real* practical project for that part featured a simple headphone amplifier, which used a complementary pair of transistors operating in push-pull emitter-follower mode. But, whereas our simple headphone amplifier was only designed to provide an output power of a few tens of mW, most practical power amplifiers need to produce at least an order of magnitude more in terms of power output – often much more. To do this, we need to use supplies that can deliver an appreciable voltage and current and also ensure that the transistors used (both drivers and output devices) are suitably rated and also fitted with heat sinks, where appropriate. We might also want to use some of the more sophisticated building blocks that we've introduced in Part 6 and 7 of the series. For example, the V_{BE} multiplier provides us with a neat way of adjusting and regulating the bias applied to the output transistors, as we shall see later.

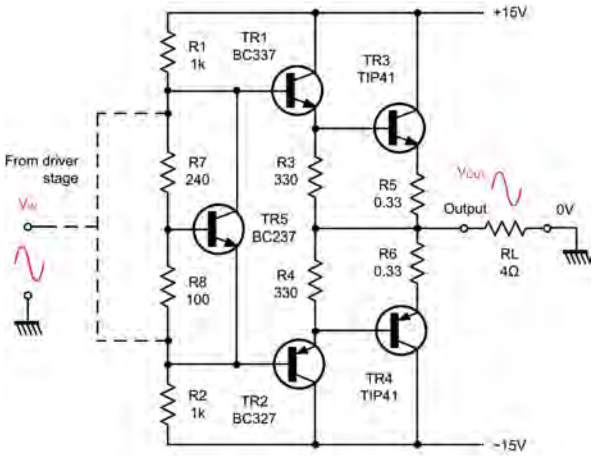


Fig.8.8 Output stage using cascaded complementary emitter followers

Cascaded emitter followers, Darlington transistors, and cascaded complementary emitter followers

When connected as an emitter follower, a bipolar transistor is typically capable of providing a current gain in the range 50 to 200. By connecting two such stages in cascade (such that the emitter of the first device supplies base current to the second device) the current gain can be raised to an exceptionally high value. Let's assume that TR1 in Fig.8.7a has a current gain (h_{FE1}) of 150 at a modest value of collector current, while TR2 in Fig.8.7a has a current gain (h_{FE2}) of 50 at the much higher value of collector current that would be prevalent in the output stage of a power amplifier. The combined current gain of the arrangement would then be:

$$h_{FE} = (h_{FE1} \times h_{FE2}) + h_{FE1} + h_{FE2}$$

If both h_{FE1} and h_{FE2} are large, then this relationship can be simplified to:

$$h_{FE} \approx h_{FE1} \times h_{FE2} = 150 \times 50 = 7{,}500$$

This implies that an input current to the base of TR1 of as little as 200µA will produce an output current of around 1.5A from the emitter of TR2. Clearly, this arrangement is going to be rather handy in the output stage of an amplifier where a significant current is required to drive a low resistance load such as a 4Ω or 8Ω loudspeaker.

The compound transistor arrangement shown in Fig.8.7b is referred to as a Darlington transistor after its inventor, Sidney Darlington. This device was originally patented in 1953 when Darlington was working at the Bell Laboratories in the USA and, since then it has provided the electronic circuit designer with a simple and elegant way of providing a large amount of current gain. Note that the arrangement in Fig.8.7a is also frequently referred to as a 'Darlington pair' but, strictly speaking, this term should be reserved for the case in which the two transistors are fabricated on the same slice of silicon.

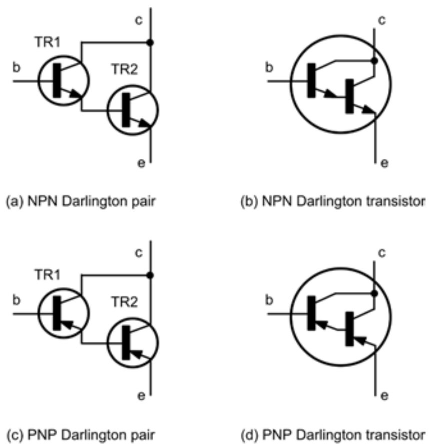


Fig.8.7 Cascade emitter follower and Darlington transistor configurations

Darlington transistors, like their individual discrete counterparts, are often available in complementary, NPN and PNP pairs as shown in Fig.8.7b and Fig.8.7d. Power Darlington transistors can also be supplied as 'gain matched pairs' but such devices tend to be rather expensive. For this reason we have avoided their use in this series.

Fig.8.8 shows a simple output stage based on cascaded complementary emitter followers (ie, a Darlington configuration based on discrete devices). Complementary NPN and PNP devices are used both in the driver stage (TR1 and TR2) and also in the final emitter-follower output stage (TR3 and TR4). The arrangement provides a voltage gain of slightly less than unity but, as mentioned earlier, it provides considerable current gain. Note how the circuit uses symmetrical positive and negative supplies and is fitted with a V_{BE} multiplier stage (TR5 – see last month) which establishes the bias current for the output stage.

Using a symmetrical ±15V supply the circuit is able to deliver an undistorted output of around 12.5W to a 4Ω load.

Table 8.2 DC bias conditions for the transistor in Fig.8.8

Device	Emitter	Base	Collector
TR1	0.539V	1.22V	15V
TR2	–0.552V	–1.24V	–15V
TR3	–15V	–14.55V	–0.047V
TR4	–1.24V	–0.513V	1.22V
TR5	0.047V	–0.522V	–14.55V

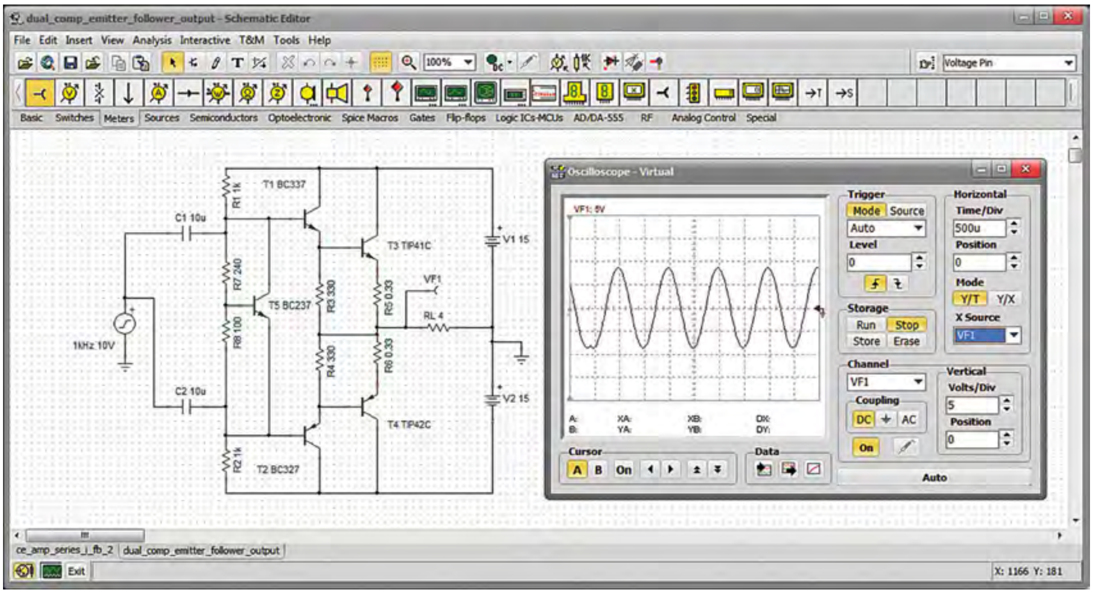


Fig.8.9 Testing the cascaded complementary emitter follower output stage

Fig.8.9 shows the circuit on-test using Tina Design suite (this circuit is available for download from the *EPE* website). Note how the signal source has been adjusted to provide an input signal of 10V peak (20V pk-pk) at 1kHz. In order to check bias conditions it is useful to perform a DC analysis of the circuit before attempting to perform a signal analysis or display waveforms using Tina's virtual oscilloscope. If you do this you should obtain the values indicated in Table 8.2. Note that you can use Tina's 'DC Table' tool in order to generate the results for your circuit all in one go!

Checking symmetry, bias and quiescent (no-signal) set-up

If you take a close look at the full DC voltage table generated by Tina Design Suite you should note that the quiescent voltage developed across RL is about –8.55mV instead of the 0V that would be expected from a perfectly balanced circuit. However, at a mere –8.55mV it is quite close!

With the values shown for R6 and R7, the quiescent current present in the output transistors (TR3 and TR4) is quite large for a Class AB amplifier at a little over 60mA. This, in turn, results in a quiescent power dissipation of around 1W in each of the output devices. This should not actually be a problem since, in a real circuit, the devices will be mounted on a correctly specified heat sink (see last month). However, by reducing the value of R7 it might actually be prudent to back-off the quiescent current to around 25mA, in which case the amplifier will still be operating in Class AB mode. You might like to try this with your own Tina model to see just what range of adjustment might be needed if R7 was to be replaced with an adjustable bias-setting component in the form of a PCB-mounting miniature pre-set variable resistor.

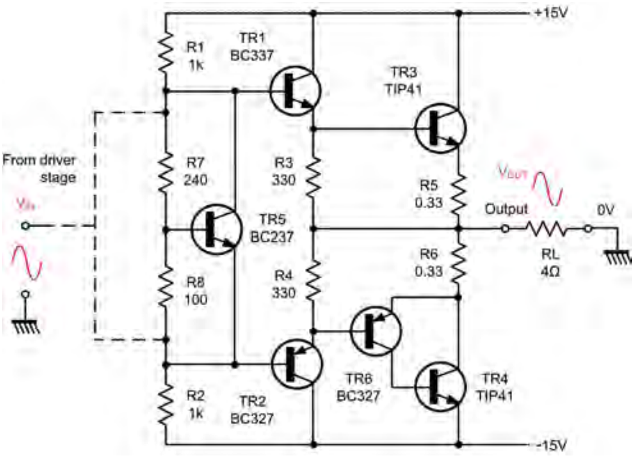


Fig.8.10 Output stage using a compound-feedback (Sziklai) pair as one of the emitter followers

The Sziklai compound feedback pair

One disadvantage of the complementary output stage is the need to provide a complementary pair of devices. In some cases the PNP device can be significantly more expensive than its NPN counterpart. For example, the ever-popular 2N3055 NPN power transistor (15A, 60V) costs around £1 in one-off quantities from one of the UK's leading suppliers, whereas its PNP3055 counterpart costs £2 from the same supplier. In addition, there is a need to ensure that the devices are reasonably well gain matched. This can be quite difficult unless suppliers are offering 'gain-matched pairs' of devices (usually at a premium). For these reasons it might be better to make use of only one type of device in the final power stage. This will invariably be an NPN device and, when selecting from a batch of similar devices there is a high probability of finding devices with very similar current gain. In fact, when selecting devices for the final practical project in this series we tested a batch of ten TIP41 NPN power transistors and found that the variation in current gain was no more than about ±5%.

In order to use two NPN devices as power emitter followers we need to add an additional device to the output, as shown in Fig.8.10. This extra driver stage is a PNP device (TR6) connected in what is known as a compound feedback or Sziklai pair configuration with TR4. The two devices, TR4

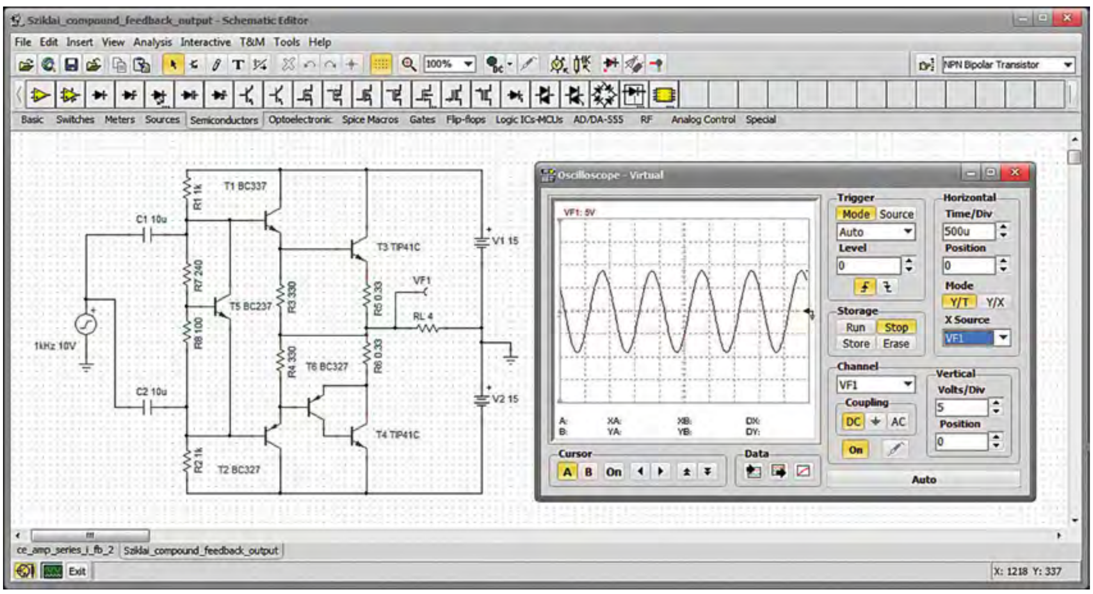
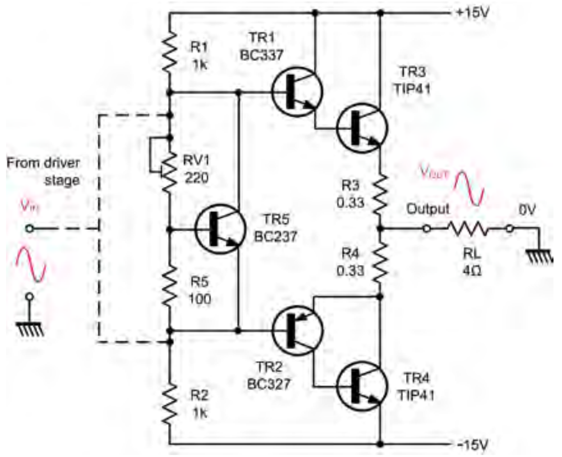


Fig.8.11 Testing the compound feedback pair output stage



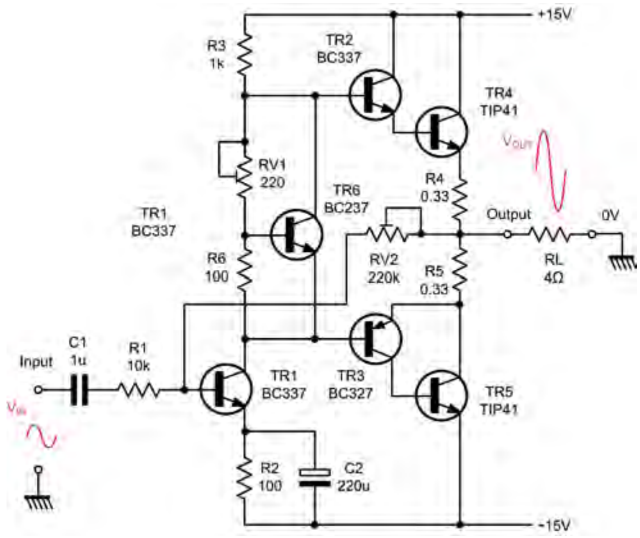


Fig.8.12 (Above) Quasi-complementary output stage
Fig.8.13 (Right) 10W quasi-complementary output with simple driver stage

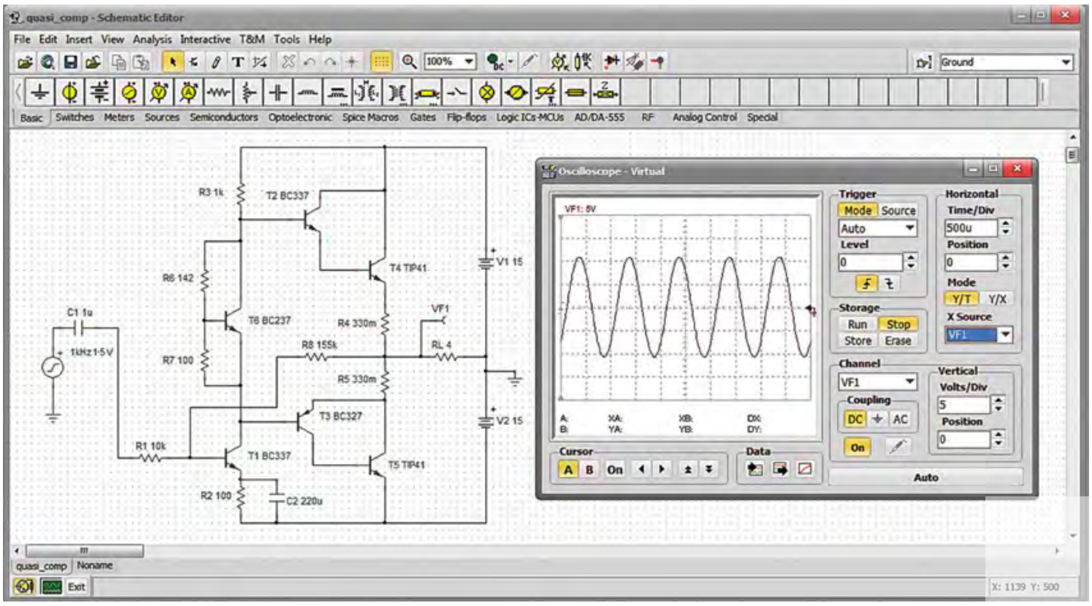


Fig.8.14 Testing the 10W simple quasi-complementary power amplifier

and TR6, behave in a similar manner to one device of opposite polarity to that of TR4, ie, they jointly replace the PNP device, TR4, in Fig.8.8. Note that the collector connection of the NPN transistor, TR4, effectively becomes the emitter connection of the PNP device that it replaces. Similarly, the emitter connection of TR4 becomes the collector connection of the replaced PNP device. This may all sound a little confusing, so we will return to this subject a little later when we examine the current flow in various different output configurations.

If you test the circuit shown in Fig.8.11 using Tina Design Suite you will find that it provides a very similar performance to that of the earlier circuit based on cascaded complementary emitter followers. Table 8.3 shows the quiescent DC voltages for the circuit. Once again, it might be worth experimenting with the value of R7 in order to produce a more modest quiescent current in the output stage, TR3 and TR4.

Table 8.3 DC bias conditions for the transistor in Fig.8.11

Device	Emitter	Base	Collector
TR1	0.539V	1.22V	15V
TR2	–0.552V	–1.24V	–15V
TR3	0.056V	0.539V	15V
TR4	–15V	–14.55V	–0.047V
TR5	–1.24V	–0.513V	1.22V
TR6	0.047V	–0.522V	–14.55V

Further increasing the output power

In order to increase the power output and to cope with low-impedance loads (4Ω or less) additional emitter followers may be connected to the output in parallel with the existing output devices. However, if this is done there is a need to ensure that the devices used are reasonably accurately matched in terms of current gain and also to ensure that sufficient current drive is available from the preceding stage. In order to assist with current-sharing and also to provide a degree of thermal protection, individual low-value emitter resistors are invariably fitted to each device. (An example of current sharing is shown later in Fig.8.16e.)

Instead of just replacing a single PNP output transistor, the compound feedback pair that we met earlier can be used in a symmetrical output stage where the NPN device is itself augmented by an additional NPN device operating in a Darlington configuration. This arrangement (see Fig.8.12) is often referred to as quasi-complementary symmetry. Note that the Darlington and compound feedback pairs are capable of exhibiting remarkably similar current gains, as demonstrated later in Fig.8.16.

Bias adjustment

In order to set the DC operating conditions for the output stage, we have included the pre-set variable resistor, RV1. This component is adjusted in order to produce a standing (quiescent) current of around 20mA to 30mA in R3 and R4. This, in turn, ensures that the output stage operates in Class AB mode with a modest standing current in order to avoid cross-over distortion that might otherwise be the case. Note that if too much standing current is present, TR3 and TR4 can dissipate an undesirably large amount of power. Note also that, with 30mA of quiescent current flowing in R3 and R4, each device will dissipate around 450mW – even with no signal present!

The circuit shown in Fig.8.12 is also lacking a driver stage and some means of balancing the output so that TR4 and TR5 share the supply voltage equally. Correct balancing of the output will ensure that, under quiescent conditions, no voltage appears across the load, RL. We've put this right in the arrangement shown in Fig.8.13.

Balancing adjustment

Balancing the output stage ensures that the supply voltage is shared equally between the two output devices and that no DC voltage appears across the load under no-signal (quiescent) conditions. Some means of adjustment is also essential in order to cope with variations in transistor current gain and in component values

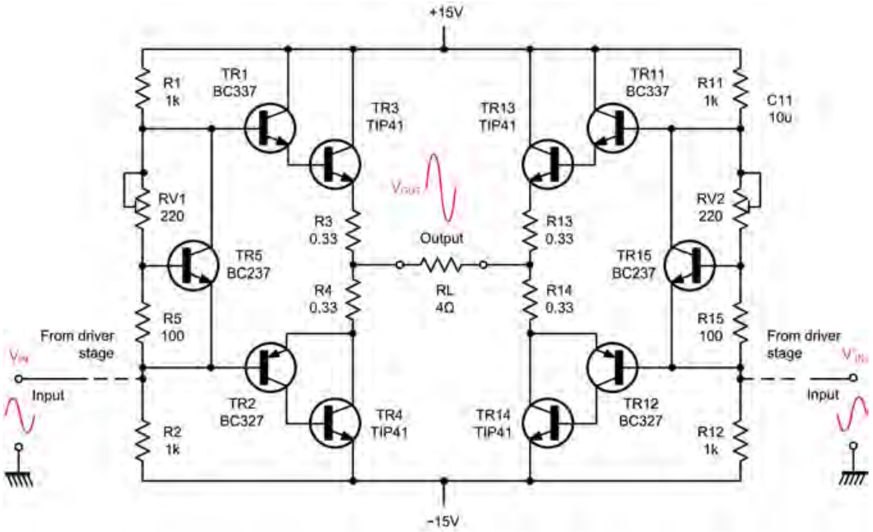


Fig.8.15 Quasi-complementary bridge-configured amplifier

where individual tolerances in resistor values may typically be of the order of ±2% to ±5%.

To balance the output stage a second pre-set resistor (RV2) has been added to the circuit, as shown in Fig.8.13. This resistor provides shunt voltage negative feedback (see earlier) and it is adjusted until the DC voltage at the output terminal is exactly 0V in the absence of an applied signal. Note that, in practice, there can be some interaction between RV1 and RV2 and so several adjustments may be required in order to establish the required operating conditions (both quiescent current in the output stage and zero output voltage).

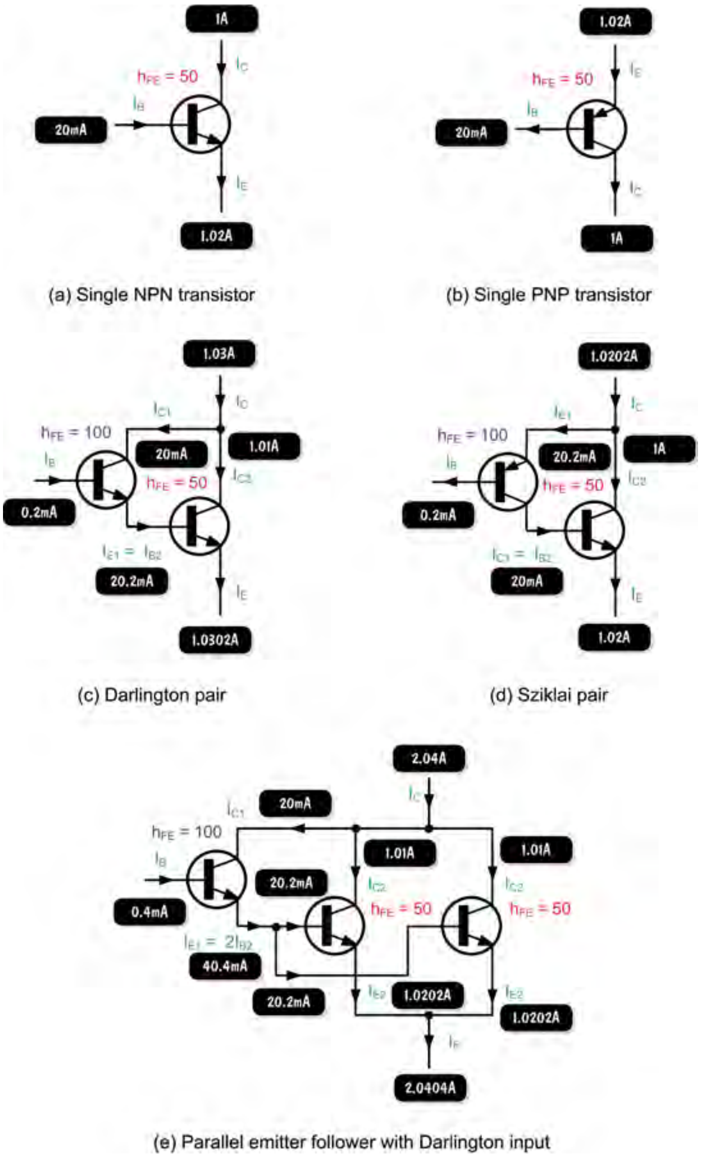


Fig.8.16 Overview of common output transistor configurations with typical peak signal currents indicated

The arrangement shown in Fig.8.13 is capable of providing an output of slightly more than 10W with only a small amount of distortion (less than 0.5%) evident. However, this circuit is still quite a long way from perfection, as we shall show next month. Fig.8.14 shows how we tested the circuit of Fig.8.13 using Tina Design Suite. In this test we have applied a signal of 1.5V peak at 1kHz to the input of the amplifier in order to produce an undistorted output of a little over 20V pk-pk (corresponding to an RMS power of 12.5W into a 4Ω load).

Bridge-configured amplifiers

A further increase in power output can be obtained (but without increasing the supply voltage) by using two power amplifier stages to drive the load with anti-phase input voltages. This arrangement is referred to as a bridge configured output stage and, in theory, such an arrangement should be able to produce a fourfold increase in output power for the same supply voltage and input signal amplitude.

The power increase results from the effective doubling of the output voltage; as one output is taken high the other output is, at the same time, taken low, and vice versa. In other words, an output of 10V pk-pk from an unbridged power amplifier would increase (in theory) to 20V pk-pk in a bridge-configured output stage. The doubling in output voltage produces a quadrupling of power in the load (recall that power is proportional to voltage squared). Unfortunately, things are rarely that simple – in order to quadruple the power in the load we would not only need to double the output voltage but, at the same time, we would need to double the amount of supply current. If the power supply is not adequately rated the supply voltage will fall and, as a result, limit the output voltage swing.

Fig.8.15 shows the circuit of a bridge-configured quasi-complementary output stage. Note that neither side of the load is connected to 0V (ground) and the voltage that appears across the load will be the difference in the output voltages from the individual quasi-complementary stages. Thus, on the positive peak of a signal having an amplitude of 10V, the voltage appearing on the load will be +10V from one output (the junction of R3 and R4) and –10V from the other output (the junction of R13 and R14), making a total of 20V. On the negative peak of the same signal, the voltage appearing on the load will be –10V from the junction of R3 and R4 and +10V from the junction of R13 and R14, making a total of –20V.

Summary of output transistor configurations

To bring this part to a conclusion, Fig.8.16 provides an overview of some of the most common output transistor configurations. We've included typical currents that might be expected when an amplifier is driven on signal peaks. Notice how the drive current varies and how it is significantly less for the cascaded emitter follower (Darlington pair) and complementary feedback pair (Sziklai pair).

Next month

In next month's *Teach-In 2015*, *Discover* will be devoted to practical aspects relating to measurement, adjustment and fault-finding on power amplifiers. *Knowledge Base* will look at stability, thermal and over-current protection. In *Get Real*, our final practical project will be devoted to a low-cost high-quality 10W power amplifier that will out-perform most of today's integrated circuit amplifiers. This module is ideal for use with our other *Get Real* projects and will help you to build a complete high-performance audio system.

NET WORK

by Alan Winstanley

Chaos theory

SOME years ago the writer had the pleasure of playing host to Thomas Scarborough, an *EPE* contributor from Cape Town famed for his *Ingenuity Unlimited* ideas and many popular *EPE* constructional projects. We were moseying around a local town and came across a hardware store with a selection of brooms, brushes, buckets and general home hardware displayed proudly on the pavement. Thomas remarked that the whole lot would probably disappear in seconds if we lived in the chaos that he saw all the time in South Africa.

The same levels of chaos seems to have been visited upon the author by British Telecom in recent months. A new construction site suddenly appeared nearby on the spot of a former farmhouse, which was demolished to make way for a new housing estate. All was fine until the site required some phone lines, when a BT Openreach engineer was seen furtively rummaging at the base of a nearby telegraph pole. True to form, my dial tone suddenly disappeared, but (strangely) I still had my fibre broadband. The weary old routine of getting it fixed will be familiar to many readers: phone BT (difficult when you have no dial tone) to report a fault, then wait the stipulated five days for a repair. You are also reminded that you will be charged nearly £130 if the problem is somewhere on your property.

Suffice to say, many weeks passed with one failed repair after another, because as fast as one fault was fixed another one appeared, losing dial tones again and then neighbouring properties entered the melee. The old telegraph pole suddenly disappeared, replaced by a much taller one. No less than six adjacent properties then suffered an interminable string of problems with their phone services. Often the repair was a last-minute rush by harassed engineers working barely within the five-day limit. BT Openreach – the contractors who repair wires – would appear from nowhere and test oscillators would be attached to my phone socket while they disappeared down the road to test for a line break. Engineers work to a rigid system of job numbers and are not a public-facing organisation. One grumpy engineer who repaired yet another missing dialtone (for a neighbour) wasn't interested in the fact that he had just knocked my broadband off altogether when I shouted to him up a telegraph pole. I had to report it and wait another five days.

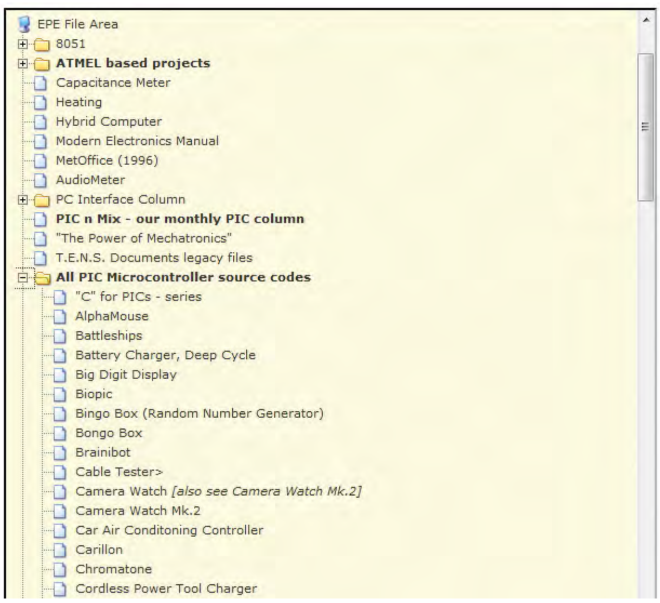
I also suffered something that I thought had long since disappeared – crossed lines. In the chaos, another property began receiving my calls but, worse, their outgoing calls were racking up on my bill (revealed by logging into my account at BT's website to check the log). After expending several working days of engineers' labour, so far the service is holding up again, but the whole image has been one of a creaking network teetering on a knife edge. To its great credit though, BT's overseas call centre has been very skilful and professional, and like the best of quality systems, once I escalated the fault they would not let matters lie until I confirmed for sure that the problem was fixed. More than once I have sung the praises of BT's overseas call centre staff, who have sometimes stuck doggedly with me for several hours to fix a fault with a customer's connection.

While BT's website is a bloated portal, it is worth BT customers logging in to check their accounts online, view bills and keep an eye on phone call usage if you are that way inclined: you never know if someone else is thieving phone calls using your own account.

Surfing the *EPE* way

This month's *Net Work* offers a long overdue review of *EPE's* own Internet presence and a reminder about the resources we can offer our readers online. *Everyday Practical Electronics* has provided a website for nearly twenty years (our first site from 1996 is preserved at **www.epemag.wimborne.co.uk/legacy/oct96www**), during which time the website has seen several redesigns as *EPE* and our marketplace have continued to evolve. Judging by some of the emails I receive (often in search of source codes), it is surprising how some readers overlook our main website altogether; it is accessible at **www.epemag.com**, although strictly speaking this is a 'front' for the site's true hosting address at **www.epe-magazine.co.uk**. Using either domain name will get you through to the same set of files, so you can use whichever URL works best for you.

The current website is updated every month by the writer, and it can be a complex job because of the need to collate a lot of information from various sources. Generally, the website is updated on or soon after the official publication date of *EPE*, which is approximately the first Thursday of the month. The home page carries a cover shot of the current issue, and clicking on it will take readers through to that issue's dedicated web page, where outline details of the issue's contents will be found. The main menu bar along the top is (we hope) simple



Legacy source codes (1996 – approx 2008) are preserved on EPEmag.net*. The FTP site is also still available*

Each issue has a dedicated web page on the EPE *website. The link for that month's source code is below the covershot*

to use and the most-used links are to the Projects and Library web pages. It's also worth knowing that a consistent format is used to manage the website's URLs. For starters, the monthly project pages are hosted in a '/proj' sub-directory at: **www.epemag.com/proj/mmyy.html** – so last month's issue is at **www.epemag.com/proj/0815.html** for example.

Very important for many readers is the link to each month's source code, and this is the most common question asked by website users. We know that readers are often eager to download the latest source code files as soon as possible, so it's useful to know that all files relating to that month's issue are rolled up into a single .zip file, the hyperlink to which is found under the cover shot shown on that month's webpage. Whether it's project source code, *Teach-In* updates or *PIC n' Mix* files, they are all zipped together and a program such as WinZip or free 7Zip will unpack them. Occasionally, updates appear later down the line and we do our best to modify the monthly webpage with more details. It's worth checking the monthly web page for any updates, corrections or late-breaking news about your latest project.

For free…

When microcontrollers arrived on the hobby scene in the 1990s, *EPE* was the first electronics magazine to give away its source codes for free, but the way we manage source codes has evolved over the years. Trying to collate a lot of legacy data for today's readership has been a challenge. Source code files from the 2007 issues (approximately) up to the present date are hosted at a dedicated address in our 'Library' folder. The source code sub-directory is at **/lib/src/mmyy.zip** with *mmyy* being the month and year of publication. Last month's source code, for example, is hosted at: **www.epemag.com/lib/src/0815.zip**. Readers who wish to check source codes can therefore shortcut the system by dialling the appropriate URL straight into their browser, or they can check the monthly pages for the link instead. It's also worth knowing that even if the website has not yet been updated for the latest issue's details, we try to release source codes a few days beforehand to help readers who want to make an early start on their latest project. Simply try downloading **…/lib/src/mmyy.zip** and if the file is there, it can be downloaded and unzipped.

Not forgetting FTP

We also know that many enthusiasts carefully curate their prized issues of *EPE* from years gone by. We have done our best to preserve the original legacy codes from the 1990s to 2008 (or so) for future reference, but they were handled online in a different manner at the time; in fact, *EPE* offered an FTP (File Transfer Protocol) site from 1996, and source code files were simply stored in named folders that tallied with the project name. Regular readers will know how we published FTP addresses for old projects and the good news is that we have managed to restore the original FTP site for future reference; anyone picking up an old copy of *EPE* should be able to access the source code using the original FTP URL.

Furthermore the author's own website at **www.epemag.net** hosts the old 'file tree' area (shown on page 46) familiar to loyal readers – this provides an Explorer-style front end to the same FTP site. As the files were sorted under their name it's necessary to scan down an alphabetical list until the project title is found. For instance, the software for John Becker's famous *PIC Toolkit TK3* is located in the tree under *All PIC Microcontroller Source Codes* then *Toolkit TK3 & Updates*, which ultimately links to: **ftp://ftp.epemag.wimborne.co.uk/pub/PICS/ToolkitTK3/**. Bear in mind that sometimes a project might have been superceded by a later design, which will be found in an altogether separate folder, so hunt around just to make sure. In most cases a plain text file in the relevant folder indicates the issue's month and year.

Readers searching for legacy files for older issues of *EPE* from the mid 1990s to roughly 2008 can also view the entire FTP site the traditional way simply by typing the URL **ftp://ftp.epemag.wimborne.co.uk/pub/** into their web browser, or use an FTP client such as Filezilla to fetch them using anonymous FTP. John Becker's free Windows software for *Teach-In 2000* will also be found there, and nostalgic readers might yearn for the Year 2000 (Y2K) patch for personal computers that will be found in the Software folder!

We now have a consistent system of file hosting in place that should serve well for the foreseeable future. The main point to remember is that more-recent source codes are readily accessible via the **epemag.com** website – simply visit the monthly page, or try the Library or Projects links along the top of our web site as alternative routes, then download the relevant zip file. For the oldest issues, readers need to check through the file tree on **epemag.net** or use the FTP address instead. For further online assistance in finding files, the place to go is Library Help at: **www.epemag.com/library-help.html**, where visitors will see a summary of useful information.

The Projects pages give a quick summary and covershot of each issue. Click through to that month's dedicated web page

Another way of skimming through *EPE* back issues is via our *Projects* page. This is simply a gallery of the year's cover shots with a link through to the corresponding month's web page. If you think you remember the cover or just need a quick reminder of the projects contained in an issue, try the Projects page accessible from the website's top menu.

A legacy of projects

It is always gratifying to receive requests from readers building older projects, sometimes for designs that were published decades ago. More recently we managed to restore the oldest *EPE* website project pages too, and these have gone online in the Projects area of our website. They can also be viewed at: **www.epemag.com/projects-legacy.html**, where you can marvel at some grainy photos, project corrections and some of our earliest online content. A good proportion of these projects would doubtless be feasible to construct today, and it's definitely worth a trip down memory lane to see if any old projects appeal to you. However, legacy designs do not always prove worthwhile to construct simply because parts may go obsolete or newer, better designs have come along since then. Nor unfortunately will we have any technical information or be able to contact the original designer or provide a reprint.

At this point we should remind readers to check that all parts are still available for an old project before commencing construction – it would be a shame to start assembly only to find some esoteric part or other is no longer available. eBay sometimes turns up trumps, and advertisers such as London-based Cricklewood Electronics (**www.cricklewoodelectronics.com**) specialise in stocking rare or unusual components, so check their website or perhaps drop them an email if you seek an unusual part for an older project.

Our online *Chat Zone* forum at **www.chatzones.co.uk** has a section entitled *Shop Talk*, which is the nearest thing to the legacy column of the same name, intended for late-breaking news and helping with project component sources. It is partially a self-help area and readers often suggest vendors of unusual parts for *EPE* projects themselves, or we will post any sourcing information that we hear of.

Visitors can also download **free reprints** of some of our most popular projects from days gone by. Simply visit the *EPE* Library and follow the link for free project PDF reprints. There are scores of popular old designs in various categories to choose from, most of them still feasible today – but again, do ensure all parts are still in stock before starting construction. The Library also plays host to .zip files related to our *Teach-In* series from *Teach-In 2000* to *2014*. Files for *Teach-In 2015*, expertly written by Mike and Richard Tooley, are currently included in each month's source code file downloadable from that month's web page, but the files will eventually appear in the Library once the series is concluded.

Within the last few weeks a new and long-overdue search engine was added to the *EPE* website, which is based on the excellent Zoom Search from Australia's Wrensoft. It replaces the previous Atomz search, a third-party service that was much-loved by many website owners but suddenly disappeared when arch-owners Adobe pulled the plug. Zoom is an immensely fast search system that is accessed via the top menu on every *EPE* web page. Owners of websites interested in adding a powerful search without running a bespoke database-driven application might check **www.wrensoft.com** for a free trial.

We hope readers will take time out to check around the *EPE* website: it's the result of a lot of hard work as we attempt to make our latest resources available online, as well as ensuring they are safely preserved for future readers to enjoy.

Queries and comments are welcome, as always to: **alan@epemag.demon.co.uk**

Audio Out special – review of Peak Analysers

If you are tired of sorting silicon wheat from chaff then Peak may have the answer for you. Jake Rothman is impressed with two of their latest semiconductor component testers – the DCA75 and Atlas ZEN50.

There are some companies whose new products are awaited with baited breath, one beginning with 'A' springs to mind. However, in this article we are not talking about mega-expensive multifunction machines without screws, where you can't change the battery, we are talking about dedicated test equipment that every electronic engineer needs.

I remember when I first started with an AVO meter with a bent needle, I could just about test diode junctions and find shorted semiconductors. In the 1980s I moved on to digital multimeters, which had an H_{fe} test function. It was always a bit of a problem finding the correct pin-out of the transistor. The meter used a different part of the socket for NPN or PNP devices – a total of 12 combinations. I developed a technique where I could find out in about five moves how to connect up the terminals correctly. In a 1995 *EPE* article, Jez Siddons designed an analyser with an LED matrix display and logic to sort through the connection combinations automatically. In 1997, the DCA50 Analyser was released; its circuitry was compactly implemented using a PIC (Fig.1. and 2).

Fig.1. Peak DCA50e semiconductor analyser from 1997 still going strong at Wilmot Audio.

Fig.2. DCA50e internal view

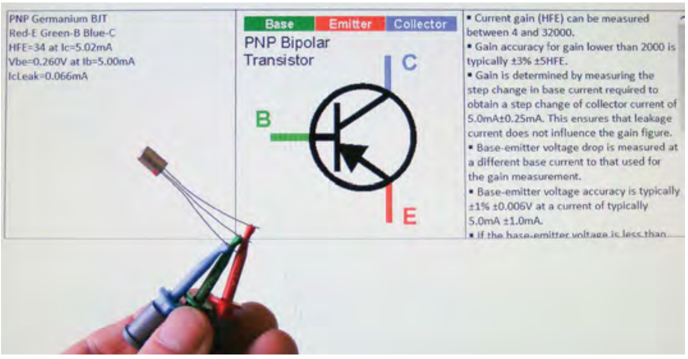


Fig.3. DCA75 testing NKT214F germanium transistor for Colorsound Tonebender fuzz box

The right kit

For any electronic engineer sorting through the usual mound of three-legged blobs, the Peak's pinout sorting ability was a godsend and this feature remains Peak's unique selling point. Later, in 1999, I started making Colorsound guitar pedals, which used infamous germanium transistors. 'Infamous' because if you tested germanium transistors on a multimeter it couldn't differentiate between leakage current

Fig.4. My old JFET tester – it Did 10 years service, but has now been replaced by the DCA75, except when I have to do a hundred of the same devices quickly.

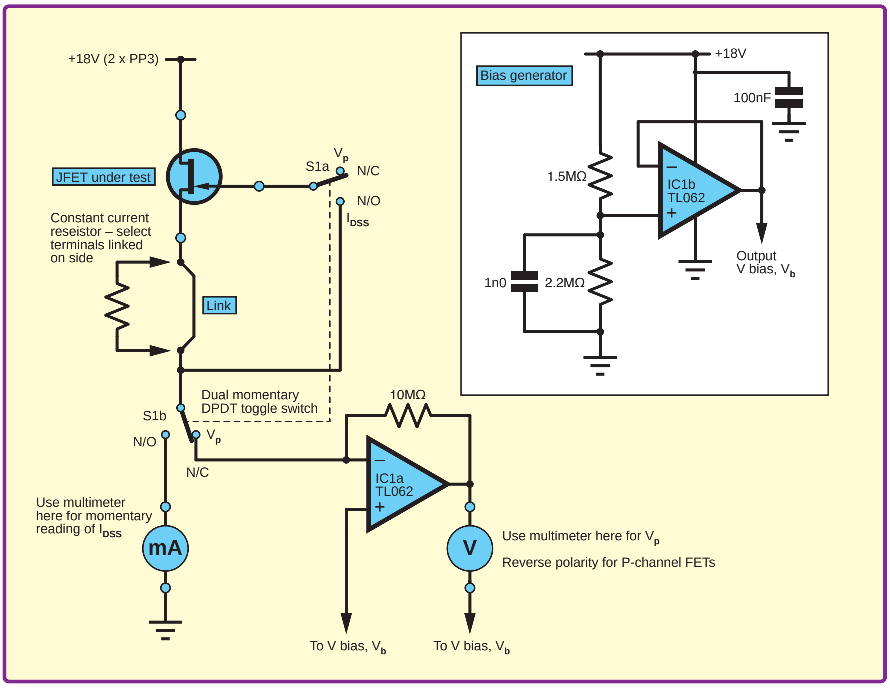


Fig.5. Circuit for JFET I_{dss} and V_p tester.

Fig.6. DCA75 readout for a JET

Fig.7. Testing a dual colour LED

and gain. That old OC71 you were testing would give a ridiculous H_{fe} of 600 or so! The earlier DCA55 component analyser sorted that problem for anyone working with such anachronistic devices. A separate reading for the leakage current is given and is removed from influencing the H_{fe} figure, and this feature is perpetuated on the new DCA75 (see Fig.3) under review here. High-quality analogue music equipment depends on component selection, since the industrial components we use are not generally optimised for audio. It's such a small market, and buying manufacturers' selected parts or those specially designed for audio is prohibitively expensive.

Another problem bedevilling the industry was the drying of electrolytic capacitors. 'Repairers' were making a living re-capping entire circuits, even if only one or two capacitors were actually faulty, and throwing hundreds of good capacitors in the bin. The Peak ESR analyser put an end to that racket with a genuine environmental benefit. Companies like Peak make a better widget, add engineering value, cut down the waste stream and employ people making real stuff. Engineering SMEs (small and medium-sized enterprises) add lasting wealth to the economy.

When the new DCA75 came out I knew it would be another bit of hardware that would enable me to do a better job. I was particularly interested in the new JFET testing function, since it could test for pinch-off voltage V_p or V_{gs} [off] and I_{dss} (the constant current value). I had designed and built a dedicated unit to do this in 2004 (shown in Fig.4 and Fig.5) Analogue guitar phasers need up to six V_p-matched FETs. Many circuits use JFETs as constant current sources, which typically have 100% tolerance, so careful selection is required. I had suggested Peak add these features in 2005, but designing test gear up to full production level takes time and it was worth the wait. I was pleased to see that our two units agreed with each other on all the JFETs I tried. Fig.6 shows the readout while testing a FET.

The unit also measures transconductance, g_{fs}, which is how the drain current varies with gate voltage. This gives an indication of the potential gain of the device, a parameter that is often too low with JFETs. I know of no other unit that can test this – so, if you are building high-impedance sensor amplifiers, say for condenser microphones, it will be essential to have this unit. I was pleased to find a load of cheap 2SK117s I bought from Tayda had a g_{fs} of 16mA/V. The best audio JFET I found was the LSK170 that had 25mA/V. The humble 2N5457 only managed 3.2mA/V.

Test gear mustn't destroy what it's testing – it's the 'tester's Hippocratic Oath'. This means when testing the I_{dss}, the current is limited to a maximum of 12mA. Measuring high-current FETs such as the J112 and U1898, which have an I_{dss} of around 50 mA, gives a reading of I_{dss} > 12mA.

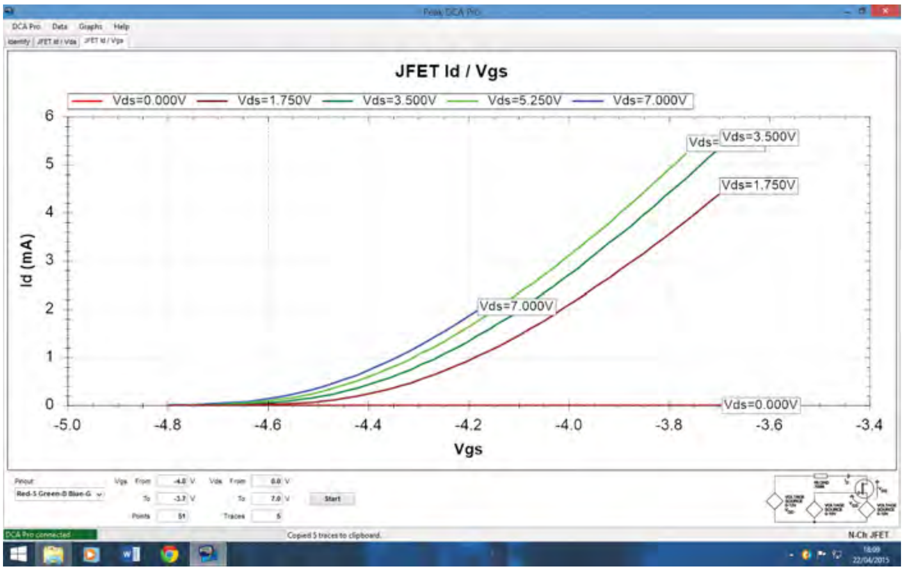


Fig.8. Curve for a U1898 JFET – note the high pinch-off voltage

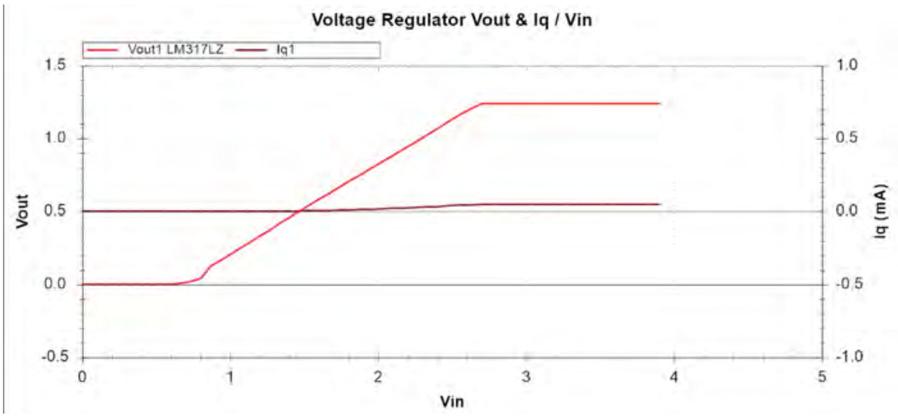


Fig.9. Curve for an LM317LZ. A 100mA adjustable regulator. From this you can see the O/P voltage, dropout voltage and quiescent current

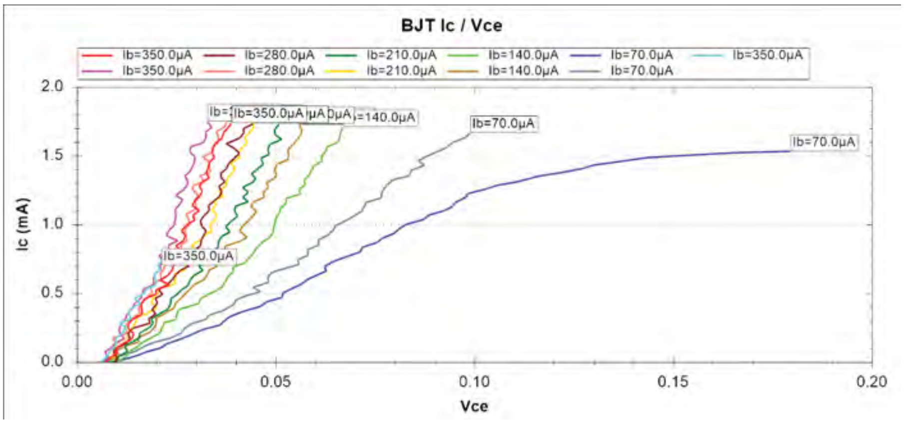


Fig.10. Attempting to match two NKT214 germanium transistors

Voltage regulators

I was impressed with the voltage regulator test function. It worked for most regulators provided they worked at 5V or less. Audio people love the variable LM317 adjustable regulators. These regulators have a tolerance of around 5% on the reference voltage, which means using the DCA75 they can easily be matched rather than trimmed. It also tells you the quiescent current (I_q) which is useful for selecting low-current units for battery operation. I was troubled by the variation between individual devices. Some small 100mA ST LM337s had a very low I_q of 0.18mA. Testing a batch of national LM317Ts, the DCA75 said 'no component detected'. I thought this was because the I_q was above the units max limit of 5mA, but it wasn't. I suspect the regulators oscillated because of a lack of decoupling capacitors, but I couldn't see any with the 'scope. I've sent some of these regulators to Jez to find out what's going on. A later software update should fix this anomaly. Another useful regulator measurement was the drop-out voltage indication. Some LM series units had less than a volt – very useful for selecting devices for battery-powered gear.

Diode networks

The unit will detect double diodes, I was impressed when I tested a dual-colour LED, as shown in Fig.7.

Curve tracing

The DCA75 can give read outs on a PC directly using the supplied USB cable and software loaded from a memory stick. The appliction loaded fine onto my Windows 8 PC. With my presbyopic eyes I find a computer display much clearer than a grey on green LCD, but such displays are still essential for long-life battery operation. I suspect this unit will find favour in education because of the graphics displaying the device symbol.

Now we come to the curve tracer function – I have an Elliot curve tracer, which I used to use for matching transistors for germanium amplifiers and synthesiser ladder filters. The Elliot was a pig to use and displayed its curves the wrong way round on the oscilloscope. In conjunction with any PC this can now be done with ease on the DCA75. A screen shot is shown in Fig.8 for a JFET. It was smoother than expected, and this is a result of using 12-bit converters and linear post regulation on the internal DC step-up switch-mode power supply. It's even possible to do a curve for a voltage regulator, as seen in Fig.9. By fiddling with the ranges I was able to compare a pair of germanium transistors, as shown in Fig.10. This would be made easier if there was a transconductance function (I_c vs V_{be}) for bipolar transistors as well as for FETs. I found that I could not print directly using ctrl-P, you have to right click on the graph and 'print' is listed there. Also, I had to right click to paste it to the clipboard, then go into Paint if I wanted to store the image as a picture. Despite these minor annoyances I spent a whole day plotting graphs of various transistors – bipolar and unipolar. It was easy to check for linearity of H_{fe} vs I_c curves and FET V_{gs} vs I_d curves.

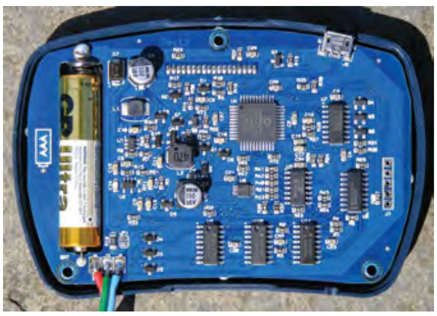


Fig.11. Inside view of DCA75 revealing conventional surface-mount asssembly

Batteries

It's so nice to see equipment powered by ordinary batteries you can get from your local shop. My AVO meter used some awful 22V thing that took two weeks to arrive. The earlier Peaks analysers used a 12V battery which was a bit of a pain, but the latest units have excellent DC-to-DC converters that use a bog-standard AAA. The units were ready to use straight out of the box with the battery already installed. Why doesn't everyone do this? I tested them for noise and my 'electromagnetic smog' detector picked up a little during the actual analysing, which then settled down to a very low quiescent level. The ZEN50 generated more noise because of its higher voltage output. It was not enough to disturb low-level noise testing on the bench. Nice to see the back is fixed with standard cross head screws, no silly internal clips to break or battery flaps to lose.

Build quality

Fig.11 shows the internals, which is an excellent example of conservative and reliable surface-mount design that should last about 20 years. The main failure modes would be the test leads falling off, battery leakage and possible chemical degradation of the LCD. There are no weird single-source components. Notice the switching matrix consisting of three FET switch chips. These are the expensive DG442, as used in good quality mixing desks. They have a low on resistance and can deal with ±12V signals, unlike the humble 4066. I notice on the website all spares are available and repair and re-calibration is cheap, although my 10-year-old units haven't drifted (the display tells me recalibration was due in Feb 2008!). The test leads have become more robust and will clip onto a TO3 case (Fig.12) without pinging off. If you are testing lots of the same device, a socket can be attached to the clip leads. A 0.1-inch Molex connector is ideal for connecting TO220 plastic power transistors quickly (Fig13.). If you are using surface mount devices, a SOT23 clamper is available, as shown in Fig.14.

Peak testers are made with lead-free solder, even though test equipment has a sensible exemption in view of its long life. However, this unit will not suffer the usual lead-free failure modes, since the board is small and suffers little mechanical flexing. It has to be done this way because the main distributors, such as Farnell and RS, only accept RoHS-compliant products since lead cross contamination could occur.

ZEN50 zener diode analyser

This is a bit of gear I've been meaning to build for years, but when I saw this for £39.00 I gave up the idea, especially since mine would have been mains powered. The unit is shown in Fig.15 and its internals in Fig.16. We've all suffered the problem of identifying strangely marked zener diodes. For example with the 1N5362, you wouldn't guess it was a 27V device and I've got loads of 1N series diodes with numbers I can't find on Google. I used to mess about with a resistor, a 60V variable bench power supply and multimeter to deal with these. Another bugbear is

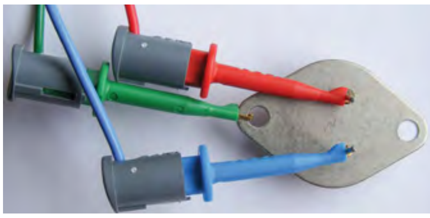


Fig.12. The new clips can attach to a TO3 case without pinging off

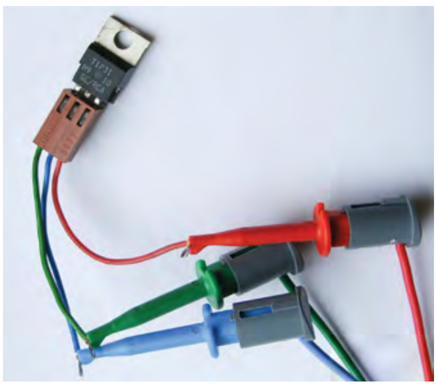


Fig.13. If you need to test transistors in quick succession a socket can be used; the 0.1-inch Molex connector is perfect for TO220 cases

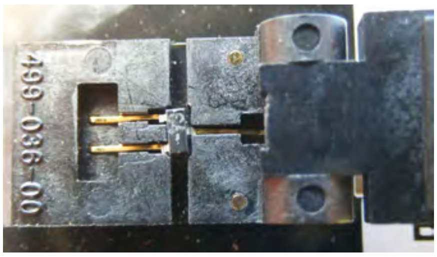


Fig.14a. DCA75 connected to Peak's PCA23 SOT23 adapter; and Fig.14b. SOT23 surface mount device in adapter prior to clamping.

those protection devices like Transzorbs and Transils (both unidirectional and directional), TVS (transient voltage suppressor) diodes and (VDRs) voltage-dependent resistors. These are now proliferating in all things switch-mode, such as LED lamp drivers. This unit enables you to sort all these devices out, so long as they operate at less than 50V. I still have to use a variac for higher voltage devices such as mains clampers.

Slope resistance

This is a unique measurement which is quite tedious to do (I remember one of those awful graph-based assessments at college). The slope resistance tells you the source resistance of the diode if you use it as a voltage reference source. This gives a guide to the regulation of the device. It confirmed one thing I had long suspected, that LEDs provide better regulation than low-voltage Zener diodes. Also, it showed that the germanium diodes popular in fuzz boxes have a high slope impedance, and this is a vital part of their simulation with silicon diodes. Since all diodes have a curved knee as they begin to conduct, it is difficult to give an exact source impedance. The Peak does a three-point analysis on the curve to give a better figure.

Fig.15. The Atlas ZEN zener diode tester

Fig.16. Atlas ZEN internals – note the large inductor to generate the required 60V

Warning

Unlike other Peak testers, the ZEN50 does not identify terminals, you have to connect the device the right way. For LEDs, you have to connect the red lead to the long lead of the device. In theory, some devices could be destroyed by almost 60V applied if the device under test is reverse biased. I did try this on some LEDs and delicate signal diodes with no adverse effects. After all, it is current limited to only 2mA and is pulsed. Peak however, have to cover themselves because someone's bound to connect an expensive antique tunnel diode, blow it up and sue. But if you really can't deal with two wires I'd advise giving up electronics!

Another useful feature is that the test current can be scrolled up. This enables the minimum current needed for stable voltage referencing to be obtained. Some zener diodes have a high slope resistance at 2mA, which then markedly reduces at 5mA. For example, a 2.7V BZY88 zener diode has a slope resistance of 160Ω at 2mA and 70Ω at 5mA. But I won't be using low-voltage zeners anymore because I found a green LED that works at 2.77V/55Ω at 2mA; three-times better! Transzorbs have a very low slope resistance, a fat 5W-sized 12V one I tested was 2Ω at 5mA.

Table 1 shows some popular diodes I tested. All at the standard 2mA test current unless stated otherwise. It kept me amused for a couple of hours!

Interestingly, if you are stuck without a zener diode you can always reverse bias a transistor base emitter junction as shown in Fig.17, which shows a BC182 giving 8.2V with a slope resistance of 15Ω. A PNP BC556 gave 10.5V, again at 15Ω. An NKT214 germanium transistor I tried did not 'zener' or did at over 50V.

So, it's time to get that workshop silicon slag heap sorted and these are just the tools to do it. I hope Peak do a power transistor analyser that checks H_{fe} at high currents in the future. Maybe I'll find some fake devices!

Fig.17. If you need a zener diode for around 7 to 10V, a transistor can be used by reverse biasing its base-emitter junction.

Table 1: Diode testing

Device	Voltage drop	Slope resist
1N4001 Standard Si	0.6V	20Ω
Above at 15mA	0.7V	2Ω
1N4148 (small Si)	0.65V	20Ω
Infra-red LED	1.08V	1.08V
Standard red LED	1.71V	40Ω
Hi bri red LED	1.67V	20Ω
Hi-Bri yellow LED	1.87V	20Ω
Standard green LED	1.93V	20Ω
Hi-bri Green LED	2.77V	55Ω
White LED	2.84V	70Ω
Blue Rapid 55-1480	3.24V	54Ω
1N60 Ge	0.46V	85Ω
CG92H Ge	0.41V	55Ω[1]
OA91 Ge	0.49V	100Ω
OA10 Ge power	0.29V	20Ω
Above at 15mA	0.38V	2Ω
1N5817 schottky	0.21V	15Ω
BAT86 schottky	0.26V	15Ω
BZV46 volt ref	1.50V	25Ω[2]
BZY88 3.3V zener	3.2V	170Ω
BZY88 5.6V zener	5.59V	20Ω
BZY88 6.2V zener	5.88V	5Ω[3]
BZY85 1.3W 18V zener	17.6V	5Ω[4]
P6KE43CA 43V Transil	42.9V	70Ω[5]

Notes

1 Reverse breakdown of this was 43V
2 Obsolete part – replace with LED?
3 This voltage typically lower slope resistance
4 Higher wattage generally means lower slope resistance
5 Bi-directional – same voltage both ways

CIRCUIT SURGERY

REGULAR CLINIC — BY IAN BELL

Noise – Part 2: analysis and calculations

LAST month, we started looking at random noise in electronic circuits; mainly from the perspective of spectra – plots of signal level over a range of frequencies. We saw that noise is typically characterised by constant or smoothly changing spectra. This is distinct from simple periodic signals (sinewave, square wave etc) which have spectra comprising discrete peaks at particular frequencies, and from complex meaningful signals, such as voice and music, which have spectra where the signal level varies in a complex way with frequency. The specific shape of a noise spectrum is often described in terms of a 'noise colour', for example white noise is constant with frequency and pink noise (also called 1/*f* noise) decreases in proportion to frequency.

As we discussed last month, in order to plot a spectrum by measurement or calculation we have to divide the frequency range of interest into a large number of small discrete bands and find the signal power in each band. If we double the width of these bands, then the power in each band (approximately) doubles. Since it is better to use a measure which is independent of the band size, power spectral density (PSD), which is measured in watts per hertz (W/Hz), is commonly used when discussing noise, and signal spectra in general. We can also use the square root of this (corresponding to the rms voltage), which is termed spectral density and is measured in volts per root hertz (V/Hz^{½} or V/√Hz) – we'll say more on this a little later.

Last month, we also looked briefly at interference and distortion, as often these imperfections in electronic systems are considered together. This month, we will concentrate more specifically on the sources of random noise from within a circuit, including how to calculate the noise produced by particular components and their contribution to circuit noise.

There are a variety of types of random noise which may be generated within electronic circuitry; these include thermal noise, shot noise, flicker noise, and avalanche noise. This noise is fundamentally due to the discrete nature of electricity at the atomic level – electric charge in circuits is carried in packets of fixed size (eg, electrons). We will look at these noise mechanisms in more detail in most of the reminder of this article, discussing some further general points about noise as we go.

Thermal noise

Thermal noise (also known as Johnson noise, Johnson-Nyquist noise, or Nyquist noise) is a fundamental property of resistors (including the internal resistances of sensors and semiconductor devices), which results in a white-noise voltage across the terminals of any resistor, even when it is not connected in a circuit. The two names refer to John Johnson who experimentally investigated resistor noise and Harry Nyquist who developed the theory. This was in the late 1920s.

Thermal noise is caused by the thermal motion of electrons. This is a fundamental property of any resistor, so whatever we do we cannot get lower noise than the thermal noise. Thermal noise cannot be reduced by improved component manufacture – however, since it is temperature-dependent reducing the temperature will reduce the noise. The thermal noise voltage generated by a resistor is given by:

$$v_{N,rms} = \sqrt{4kTR\Delta f}$$

Here, k is a physical constant known as Boltzmann's constant (1.38×10^{-23}JK^{-1} (joules per kelvin)); T is the temperature in kelvin (K); and R is the resistance in ohms (Ω). Δf is the bandwidth of interest in hertz (Hz) (ie, the range of frequencies over which we are measuring the noise). Δf is pronounced 'delta f', the delta symbol means 'change in' and so Δf represents a range of frequencies. The symbol B for bandwidth is also used instead of Δf, so the equation is then written as:

$$v_{N,rms} = \sqrt{4kTRB}$$

When analysing circuit noise we can model a real resistor (which generates noise) as a noiseless resistor in series with a voltage source producing the noise voltage given by the above equation (see Fig.1). We may also model the noise as a parallel current source if that is more convenient.

Random

The noise voltage or current is an average rms value (an rms AC voltage or current is equivalent to the DC value which would produce the same power dissipation in a resistor). Since the noise voltage is an average, it does not tell us what the noise voltage might be at any given instant in time. Noise is random, so we cannot predict the instantaneous voltage, but we may be able to say something about the probability that the instantaneous voltage is above or below a certain value, or within a given range (as mentioned last month, random noise can be analysed in statistical terms). In the case of thermal noise, such analysis shows that the instantaneous peak noise voltage will be less than five-times the rms (average) value for 99% of the time.

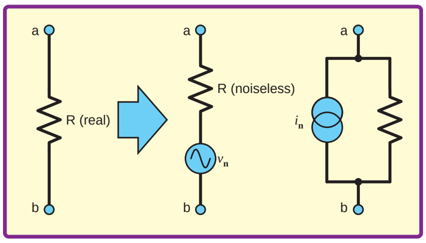


Fig.1. Resistor noise is represented by a noiseless resistor with a noise voltage source or current source

An important aspect of this equation and random noise in general, is that you have to specify a bandwidth to find a noise voltage (or current). Usually, this is equal to the bandwidth of the circuit you are using, or that of the signal of interest. However, as the cut-off of circuits at their frequency limits is not infinitely sharp, the effective noise bandwidth may actually be larger than the circuit's bandwidth. This can be characterised by finding the Equivalent Noise Bandwith (ENBW) of the circuit – the bandwidth of a circuit with infinitely fast cut-off and the same noise output.

If the specific application is not known then the bandwidth will be unknown. Therefore, noise characteristics are often expressed in 'volts per bandwidth unit' form rather than simply as voltages (have a look on data sheets for ICs such as op amps, and you will often see noise figures expressed this way). We are back to the concept of noise density, previously discussed in the context of spectra. If we divide both sides of the above equation by the square root of the bandwidth (√Δf) we get

$$\frac{v_{N,rms}}{\sqrt{\Delta f}} = \sqrt{4kTR}$$

The value $\sqrt{(4kTR)}$ has units 'volts per root hertz' often written as V/Hz^{½} or V/√Hz. So, for example, the noise from a 1kΩ resistor at 27°C (300K) is: $\sqrt{(4 \times 1.38 \times 10^{-23} \times 300 \times 1 \times 10^3)}$ = 4.07nV/Hz^{½}.

If we were interested in a bandwidth of say 20kHz the thermal noise voltage from this resistor would be 4.07nV × $\sqrt{(20 \times 10^3)}$ = 576nV

Noise temperature

Given that we know that the power dissipated in a resistor, R, with an applied rms voltage of v is v^2/R we can rearrange the thermal voltage equation as:

$$P = \frac{v^2}{R} = 4kT\Delta f$$

This applies to the situation which occurs if we short the two ends (a and b) of the noise voltage equivalent circuit in Fig.1. This scenario, where we just have a single resistor, is not typical of a real circuit or system. If the resistor is connected to another equal-valued resistor we have a situation like that when we connect something to a matched load. This is shown in Fig.2.

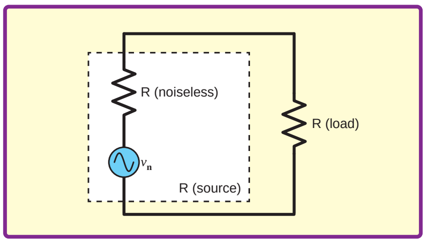


Fig.2. Noisy resistor connected to a matched load

The potential divider effect means that half the noise voltage from our resistor of interest will appear across each resistor, so the voltage in each case is:

$$v = \frac{v_{N,rms}}{2} = \frac{\sqrt{4kTR\Delta f}}{2}$$

Thus, squaring both sides:

$$v^2 = \frac{4kTR\Delta f}{4} = kTR\Delta f$$

So, the noise power transferred to the load is:

$$P = \frac{v^2}{R} = kT\Delta f$$

The power spectral density in the load due to our noisy resistor is:

$$\frac{P}{\Delta f} = kT$$

The fact that the only variable determining the PSD here is temperature leads to the idea of noise temperature, which can be used to describe the noise from a component or signal source. It does not refer to a physical temperature, but to the temperature a resistor matched to the load would have to be to deliver the same noise PSD to the system as the component or signal source of interest. The concept of noise temperature is often used in communication systems design.

Shot noise

When current flows it can generate additional white noise above the thermal noise due to the quantum nature of electric current at the atomic level. Remember, the electric current past a point is the flow of *discrete* charge carries (eg, electrons), not a purely smooth flow. This noise is known as shot noise and, like thermal noise, is due to fundamental physics and cannot be reduced. Shot noise is an important source of noise in semiconductor devices such as diodes and transistors.

Shot noise occurs when the discrete charged particles (moving due to current flow) arrive in an independent way at some point in the device/circuit. The arrivals have a specific average rate, but the actual rate varies randomly over time. Statistically, this is similar to a number of everyday situations such as the number of people entering a shop, and the number of telephone calls received by a call centre, and is known as a 'Poisson process'.

Shot noise occurs when current flows according to a Poisson process, which requires that each charged particle is moving independently. This is the case in semiconductors (PN junctions). In such cases the power spectral density of shot noise is given by:

$$2eI$$

where I is the (average) applied current in amps (A) and e is the electronic charge (charge on one electron = 1.6×10^{-19}C (coulombs)). Note that shot noise does not depend on temperature.

In metal conductors (thought of as resistors in this context) the physics is different and shot noise does not occur; the conduction mechanism smooths out the random fluctuation in the electron movement (but thermal noise, which is not related to current, still occurs). At least that is the situation for 'long' wires. At lengths in approximate range of tens of nanometres to tens of micrometres (referred to as the mesoscopic scale) metal conductors do exhibit shot noise, but at a lower level than in semiconductors.

As with thermal noise, we may be interested in the noise in terms of current or voltage and we can find this if we know the bandwidth of concern by taking the square root of the noise power. The noise current due to shot noise is therefore:

$$I_{N,rms} = \sqrt{2eI\Delta f}$$

For an applied current of 1μA this is 0.57pA/Hz^{½}, which about 80pA over a 20kHz bandwidth.

Flicker noise

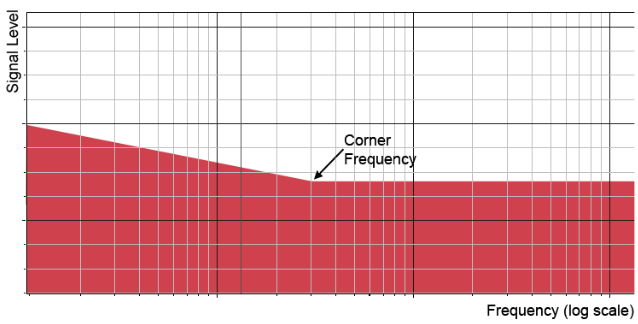
In addition to thermal noise and shot noise, resistors and active devices produce yet more noise for various and often-complex reasons. This noise is called 'flicker noise' and takes the form of $1/f$ noise, typically having a $1/f$, or similar, relationship with frequency. Unlike thermal and shot noise, the flicker noise in resistors (and other devices) depends on the component type and manufacture, and can even vary quite widely for components of the same type. For a decade bandwidth (frequency range of 1 to 10 times) the flicker noise for typical resistors varies from tens of nanovolts to a few microvolts, depending on type and quality. Carbon composition resistors produce the highest flicker noise and wirewound resistors the lowest.

The typical spectrum of noise output from a wideband circuit such as an amplifier is shown in Fig.3. At low frequencies the $1/f$ noise dominates. At higher frequencies the white noise due to a combination of thermal and shot noise dominates (this is also referred to as flatband noise, as it does not vary with frequency). The frequency at which the dominant noise changes is the corner frequency. At higher temperatures the flatband noise is mainly due to thermal noise, but at lower temperature shot noise becomes more significant.

Avalanche noise

Avalanche noise is produced by Zener diodes (or other diode junctions undergoing Zener or avalanche reverse breakdown). Avalanche noise is much larger than shot noise and so Zener diodes can introduce a lot of noise into a circuit. For this reason they should be avoided in low-noise circuits, even

Fig.3. Typical noise spectrum of an amplifier – flicker noise dominates at low frequencies and has a 1/f characteristic. At frequencies above the corner frequency, flatband thermal and shot noise dominate



though they are a temptingly easy way to produce a stable voltage reference. On the other hand, this may be useful if what you need is a source of noise.

Analysing noise in circuits

We modelled the noise from a resistor as an ideal resistor plus a voltage source producing the noise. The same approach can be used with any circuit, such as an amplifier. We replace the noisy circuit with a noise-free version and add a voltage source to represent the noise. We can put this noise-voltage source at either the input or the output of the circuit – this is termed input-referred noise and output-referred noise respectively (see Fig.4).

The values of the input and output noise are related by the circuit's gain with respect to the input noise source. In a similar way, the contribution of any individual component in the circuit can also be represented by an input-referred or output referred value.

Simple circuit example

The circuit in Fig.5, which is a potential divider, can be used to demonstrate noise analysis. We will assume the voltage source is noise free (it is an ideal voltage source as far as the noise analysis is concerned), so the noise contributions to the output will be due to the two resistors. Using the value from the calculation we did earlier, each resistor will produce a noise density of $4.07\text{nV/Hz}^{½}$ at a temperature of 27°C.

Noise calculations have to take account of the effective circuit gain from each component to the output. For the circuit in Fig.5, the gain with respect to the noise from each resistor is 0.5, due to the potential divider effect of the two equal resistors. To see how this works, draw an equivalent circuit for each resistor in turn, based on Fig.1, setting v_s to zero (removing it) so that you are just looking at the noise contribution of the one resistor. Thus, each resistor contributes $0.5 \times 4.07\text{nV/Hz}^{½} = 2.035\text{nV/Hz}^{½}$ of noise to the output signal.

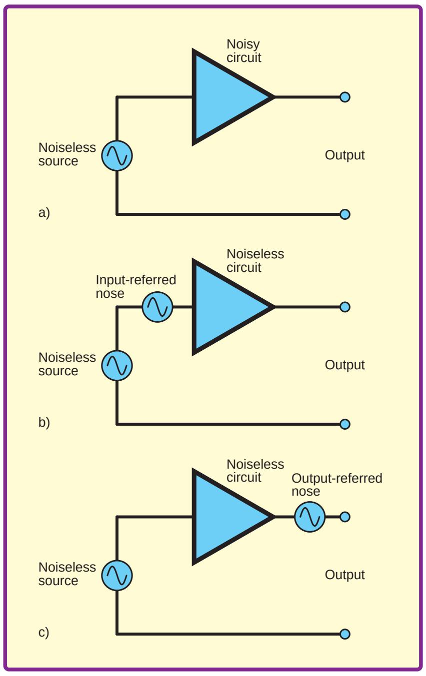


Fig.4. Representing noise a) noisy circuit driven by noiseless signal source, b) circuit noise represented by an input-referred noise source driving a noiseless version of the circuit, c) similarly, output referred noise

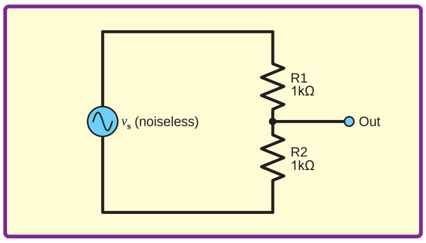


Fig.5. Simple circuit to illustrate noise analysis

To calculate the total noise density due to all components in the circuit, the square of the noise density is multiplied by the square of the gain relevant to each component, and these values are totalled (squared because this is a power-based calculation). This gives the square of the output-referred total noise density – taking the square root gives the noise density. As with individual components, dividing the output-referred noise density by the input-to-output gain of the circuit gives the input-referred value.

As resistor noise is equal at all frequencies, and the gain from both R1 and R2 to the output is 0.5, the total output noise density for the circuit in Fig.5 due to the two resistors contributions of 4.07nV/Hz½ will be: $\sqrt{((0.5)^2 \times (4.07 \times 10^9)^2 + (0.5)^2 \times (4.07 \times 10^9)^2)}$, which is $2.88\text{nV/Hz}^{½}$ at all frequencies (output referred). The input to output gain (from in to out in Fig.5) is 0.5, so the input referred noise is $2.88/0.5 = 5.76\text{nV/Hz}^{½}$ at all frequencies.

Op amp example

We will use the circuit in Fig.6 to illustrate a more complex noise calculation by estimating the signal-to-noise ratio of at the output, assuming a 1V rms signal output. To keep things manageable we will assume a noiseless input and only consider thermal noise from the resistors plus the op amp's input voltage noise. For the latter, we will use a figure of 2nV/√Hz (this value would be available from the datasheet of a real device, and in this case would make the op amp a low-noise device).

We will assume signals of interest are in the audio-frequency range 20Hz to 20kHz and that the temperature of the resistors is 25°C (298K). We will further assume that the corner frequency is below 20Hz so that we are only dealing with flatband noise.

The circuit in Fig.6 is a standard non-inverting op amp amplifier. The gain from signal input to output is, according to the usual formula for this circuit, 1 + R2/R1 = 101. When analysing noise, remember that we need to consider all components and that the gain to the output with respect to various components may not be the same as the signal gain from the input.

Op amp noise specifications usually give both a noise voltage (density) and noise current (density) referred to the input. Here, for simplicity, we will only use the input voltage noise. The op amp's own input voltage noise is always amplified by the non-inverting gain, even if the circuit is configured as an inverting amplifier as far as the signal is concerned. For this reason the non-inverting gain (in any configuration) is often referred to as the noise gain.

In this case $2\text{nV/Hz}^{½}$ of op amp noise is multiplied by the noise gain (1 + R2/R1) = 101 to give $202\text{nV/Hz}^{½}$ of output-referred noise density.

The Johnson noise density from the 10MΩ resistor is $\sqrt{(4 \times 1.38 \times 10^{-23} \times 298 \times 1.0 \times 10^7)} = 405.6\text{nV/Hz}^{½}$. This is directly connected to the output (so the gain is 1), giving $405.6\text{nV/Hz}^{½}$ output referred noise density.

The Johnson noise from the 100kΩ resistor is $\sqrt{(4 \times 1.38{\times}10^{-23} \times 298 \times 1.0 \times 10^5)} = 40.56\text{nV/Hz}^{½}$. Because this resistor is connected to the inverting input this is multiplied by the inverting gain (R2/R1) = 100, giving $4056\text{nV/Hz}^{½}$ output referred noise. Although the circuit design is of a non-inverting amplifier with respect to the signal input, it behaves as an inverting amplifier with respect to voltages applied in series with the 100kΩ resistor.

To find the total noise at the output we sum the squares and find the root. So the total output noise density is $\sqrt{(202^2 + 405.6^2 + 4056^2)} = 4081\text{nV/Hz}^{½}$ output referred noise. Over the bandwidth of interest the output noise voltage is $4081\text{nV} \times \sqrt{(20000 - 20)} = 574\mu\text{V}$ rms.

The signal-to-noise ratio in our example is therefore $20\log(1/5.74 \times 10^{-4}) = 64.8\text{dB}$. Here the output noise is dominated by the thermal noise from the resistors. Much lower resistor values could be used without affecting circuit performance and this could greatly improve the signal-to-noise ratio. In this case, the resistors are not connected directly to the signal source – the resistor ratio R2/R1 sets the gain, so lower values of both resistors can be used. In cases, where resistors are more directly in the signal path (eg, for matching purposes) it may not be so straightforward to improve noise by changing their value.

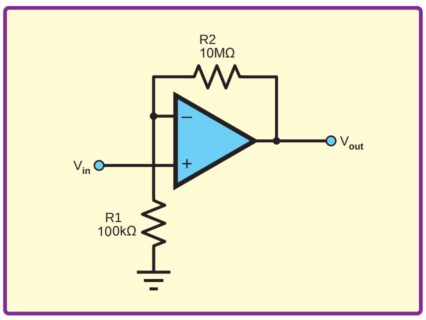


Fig.6. Op amp circuit for noise analysis example

Practically Speaking

Robert Penfold looks at the Techniques of Actually Doing it!

Integrated circuit terminology

ELECTRONICS, as with just about anything these days, has an abundance of jargon. Some of these are the most technical of technical terms, while others are simpler. The fad for acronyms has inevitably led to many electronic terms being reduced to a few letters. For beginners, even the simplest of technical names can be a bit confusing initially. Here we will consider some of the technical descriptors that are likely to be encountered when dealing with integrated circuits, or 'ICs' as they are usually called.

Integrated jargon

ICs seem to be surrounded by more than their fair share of technical jargon, together with accompanying acronyms. These terms and acronyms can to a large extent be divided into two categories – those concerned with the *physical* attributes of the components, and those that describe their *electrical* characteristics. Starting with the physical aspect of ICs, the term 'DIL' (dual in-line) is one that will be encountered very frequently. The ICs used in projects for home construction are mostly of the DIL variety, and they have two rows of pins. Fig.1 shows some DIL-style ICs in a variety of sizes. The pin spacing in each row is 2.54mm (0.1-inch). The row spacing is usually 7.62mm (0.3-inch) for devices having up to 24 pins, and 15.24mm (0.6-inch) for devices that have upwards of 28 pins. However, ICs having 22, 24, and 28 pins can have the narrower row spacing *or* the wider type.

'DIP' (dual in-line package) and 'DIPP' (dual in-line pin package) are alternative names for DIL. Components other than ICs that have DIL packages include banks of miniature switches and LED displays. These are usually termed DIP or DIPP components rather than DIL types. The three terms all mean the same thing though, and can be used to describe any component that has two rows of pins. There are also 'SIL' (single in-line), 'SIP' (single in-line package), and 'SIPP' (single in-line pin package) components. These are relatively rare, but are sometimes used for simple LED displays and banks of resistors. Again, the three terms provide different ways of referring to the same thing. Last and definitely least, there used to be 'QIL' (quad in-line) ICs. These were similar to DIL devices, but the pins in each row were arranged in a zigzag arrangement giving what was effectively four rather than two rows of pins. This type of package became obsolete some years ago.

On the surface

At one time it was common for ICs to be available in a variety of case styles, but they were all basically the same. Some had plastic cases while others had ceramic encapsulations, but they were all DIL types. The situation is different these days, with many devices being made with three or four genuinely different case styles. In addition to the DIL device, there will usually be at least one surface-mount device (SMD) version.

Conventionally, components have their leads or pins fitted through holes in the PCB (printed circuit board), and they are then soldered to copper pads on the other side of the board. With SMDs there are no holes in the circuit board, and the components are soldered to the same side as the pads. The ends of the pins are bent outwards, parallel to the PCB surface so that they fit flush onto the pads and can be soldered to them (Fig.2). This method was designed to make mass production easier, and in the normal scheme of things components are glued in place by robots, and then soldered *en masse* by a machine that produces waves of solder.

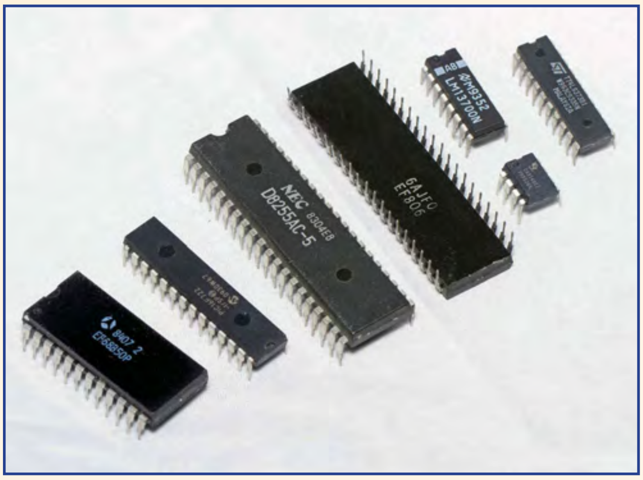


Fig.1. DIL ICs with 8 to 40 pins. They all have the pins in each row on a 2.54mm (0.1-inch) pitch. The rows of pins can be 7.62mm (0.3-inch) or 15.24mm (0.6-inch) apart

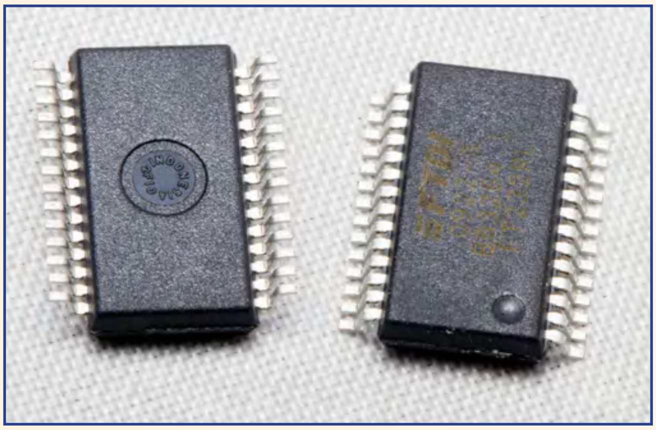


Fig.2. ‘Insect-like’ surface-mount chips do not have pins, but instead have what are more like short legs and feet. The ‘feet’ are soldered to copper pads on the circuit board

The surface-mount method of construction is not popular with electronic project constructors, those developing commercial designs, or practically anyone producing electronic gadgets by hand. In principle, surface mount construction is no more difficult that using a conventional circuit board. Indeed, it is in some ways simpler than the conventional approach, and a version of it was the preferred method for some constructors long before automated surface-mount production came along. However, the problem with surface-mount construction in a modern context is that the components are tiny in comparison to their conventional equivalents. Fig.3 shows a circuit board that has an 8-pin DIL chip on the right, and a large surface-mount chip on the left. The pin spacing of surface mount chips is not 2.54mm, and at most it is only half that figure. It can be as little as 0.5mm. Soldering a surface-mount chip in place by hand is extremely tricky, and this type of component is *not* designed to be hand soldered.

Many surface mount chips are of the ‘SOP’ (small outline package) variety, and these are also known as ‘SOIC’ (small outline IC) devices. There are variations on this encapsulation, such as the even smaller ‘MSOP’ (mini small outline package) type. The more complex surface-mount chips have many pins that are fitted in square cases with pins on all four sides, as in the example of Fig.3.

When buying ICs it is clearly important to ensure that you do not accidentally buy a surface-mount type when an ordinary DIL component is required. IC type numbers usually include a one- or two-letter suffix that indicates the type of casing. However, manufacturers often fail to use properly standardised code letters, and some chips are produced by more than one maker. Consequently, two chips of the same type could have different suffixes, but the same case style. Where there is any doubt, it is best to carefully check the descriptions in the project article and in the retailer’s catalogue to ensure that they properly match up.

Fig.3. This circuit board has a DIL IC on the right, and an SMD on the left. Surface-mount ICs are intended for automated construction methods – soldering one by hand can be difficult

Disappearing act

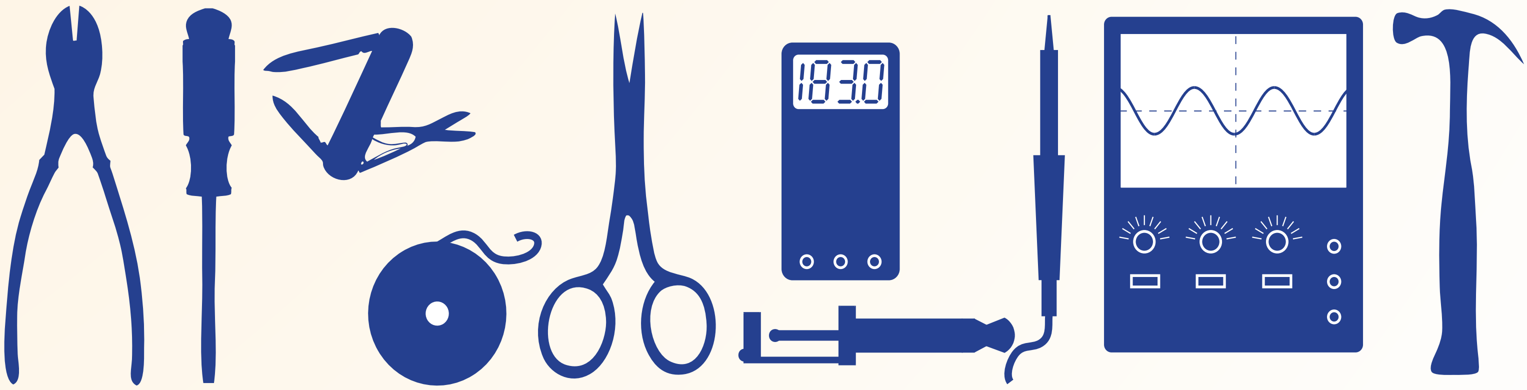
Unfortunately, it is not unknown for chips that were once available in DIL and surface-mount versions to have the DIL type discontinued after a year or so. The DIL versions are initially popular because they are needed by the design engineers who develop products that use the new chips. Thereafter, it is the SMD versions that the production engineers use. Demand for the DIL chips then dwindles, and they are no longer manufactured.

The fact that the DIL version of a chip has gone out of production does not necessarily mean that any projects that require it can no longer be built. There are sometimes stocks of components available from some suppliers long after they have been discontinued by the manufacturer. It is certainly worthwhile checking the Internet to see if they can be obtained.

Adapters provide another possible solution to the problem. The DIL and SOP versions of chips normally have the same pin functions. This makes it relatively easy to produce an adapter that enables an SOP device to be used with a PCB designed for the DIL version. Some adapters are very expensive, but these are intended for repeated use when developing prototype equipment. At its most basic, one of these devices just consists of a small printed circuit board fitted with two rows of pins. The surface-mount chip is soldered to the circuit board, which is likely to be tricky, but you then have what is effectively a DIL version of the chip. Although the cost of a basic adapter is usually quite modest, it could still cost more than the chip it is adapting, but it is usually the most practical solution to the problem.

The chop

The characteristic that separates ICs from other semiconductor devices is that they have multiple components on the same chip of semiconductor material (usually silicon). Transistors are usually manufactured by forming large numbers of them on a thin disc of silicon, which is then cut to produce the individual transistors. With ICs, things are taken a step further before the slice of silicon is ‘given the chop’. Other components are formed on the slice of silicon and further processing is used to provide interconnections that produce a complete circuit on each chip.



This integrated approach to things has its origins in the early 1950s, but it took some years to solve the problems involved when putting the basic idea into practice. In particular, there was the problem of avoiding unwanted interconnections, and 'parasitic transistors', which are transistors that inadvertently formed as part of the manufacturing process. Even with modern chips there can be problems with parasitic transistors if one of the input pins is taken slightly outside its intended voltage range. An affected device could simply malfunction, but there is also a risk of large currents flowing and damage occurring. Terms such as 'thin film technology' and 'thick film technology' refer to the particular system used to make working ICs.

The early ICs were quite simple and in some cases had less than a dozen components. I remember using early devices that were the equivalent of a very few discrete components, and the novelty of using an IC was the main reason for using them. The circuits became more complex as the technology progressed, and their relatively small size and low cost gave them real advantages over discrete circuits. Simple chips are termed SSI (small-scale integration) devices, and the slightly more complex types are called MSI (medium-scale integration) chips. At the opposite end of the range, modern microprocessors and memory chips are known as ULSI (ultra-large-scale integration) devices. These contain many millions of transistors per chip. In between there are LSI (large-scale integration) and VLSI (very large-scale integration types).

These terms are largely of academic interest, as is SOC (system on a chip). An SOC IC is a complete computer (or other complex system) which has all the active components on a single chip. There will usually be a few discrete capacitors, a crystal, and other passive components, plus some active and passive components in the power supply. There might even be an additional chip to boost the capabilities of the SOC, as in the two-chip Raspberry Pi computer.

Linear and digital

Many of the terms and acronyms encountered when dealing with ICs describe their electrical characteristics. ICs tend to be divided into two broad categories – linear and digital. Logic or computer chips are digital: everything from simple logic gates to microprocessors and multi-gigabyte memory chips are classed as digital devices. Specialised chips such as digital clocks and sound generators are also classed as digital devices. If a chip deals in signals that are 'high and low' logic levels, then it is a digital type, regardless of its function.

The linear category covers everything else, including audio amplifiers and processors, radio chips, voltage regulators, temperature sensors… and many other types. If a chip deals in varying voltages rather than logic levels, then it is linear, whatever its function. Of course, there are numerous chips that have a mixture of linear and logic circuitry. An analogue-to-digital converter takes in a varying voltage and converts this into a stream of equivalent digital values. These days many sensors have a built-in converter and provide the output value in digital form rather than as a voltage level. Devices of this type are usually categorised as digital chips, presumably on the basis that in use they form part of a computer-based system.

Logic ICs are mostly of the TTL (transistor-transistor logic) or CMOS (complementary metal-oxide semiconductor) varieties. 'CMOS' refers to the type of technology used in production – they are based on field-effect transistors (FETs). The original 4000 series of CMOS devices are relatively slow, but are adequate for many purposes, and draw very low supply currents when static or operating at low frequencies. They can also operate over a wide supply voltage range.

Other forms of MOS technology are often used for digital devices, such as microprocessor support chips. All MOS devices have extremely high input resistances, and this makes them especially vulnerable to damage from static charges. It is important to observe standard anti-static handling precautions when dealing with MOS devices.

Mix and mismatch

TTL devices are based on bipolar junction semiconductor technology, and the name refers to the type of internal circuit used. In the early days of logic ICs there were alternatives, such as DTL (diode-transistor logic) and ECL (emitter-coupled logic) chips. The TTL devices with the well-known '74…' type numbers were the ones that endured, and they are still in use today, but only in various improved versions. One of these is the low-power Schottky type, that have 74LS series type numbers. The Schottky diode enables these devices to operate at lower supply currents and higher frequencies than the standard/original TTL types.

The 74HC and 74HCT series are more recent developments, and they are based on a refined CMOS technology. They provide the speed of the 74LS family with the advantages of CMOS technology, such as minute current consumption at low operating frequencies. 'HC' stand for high-speed CMOS. The 74HCT devices are based on the same technology, but are designed to operate at TTL rather than CMOS logic levels. The 74C family are pin-for-pin compatible to their equivalents in the standard 74 series, but they use the standard CMOS technology, and their electrical characteristics are the same as the 4000 series. They are really an extension of the 4000 series rather than true TTL chips. There have been several other families of improved TTL logic devices, but none of them are still in use to a significant degree.

On the face of it, there should be no problem if you use (say) a 74LS32 chip instead of a 74HC32 type. However, although these two devices are pin-for-pin equivalents, they have different electrical characteristics. You also have to bear in mind that the supply voltage ranges of the TTL families are different, with the 74LS chips only being intended for 5V operation, while the 74HC and 74HCT chips can operate with supplies over the 2-6V range. ICs from different TTL families will sometimes work properly together, especially when a 5V supply is used, but there is no guarantee of success. It is something that is best avoided as far as possible.

ZIF

DIL ICs are normally fitted to the circuit board via a simple holder. This avoids heat damage caused by soldering the chip in place and reduces the risk of static damage. With a normal DIL socket the IC is simply pushed into place. If you need to remove it, the chip is gently prized free, preferably using a special removal tool so the pins are not damaged.

A ZIF holder (Fig.4) is more complex and expensive, but it enable ICs to be easily plugged in and removed, with no risk of physical damage occurring. ZIF stands for 'Zero Insertion Force'. The socket has a small lever, and with this raised the chip can simply be dropped into place. Lowering the lever locks the chip in place, but it can be easily removed again by raising the lever. ZIF sockets are used for things like programmers where it is necessary to fit and remove numerous devices. They are also use for large and expensive chips, such as microprocessors.

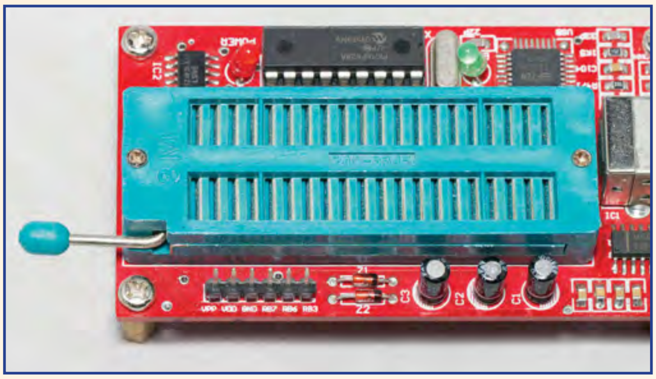


Fig.4. A ZIF socket on a PIC programmer. The lever is raised so that the IC can be fitted, and then lowered to lock it in place

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This is almost a completely opposite system to the Raspberry Pi. We learn to use a relatively simple bare microcontroller. We make our connections directly to the input and output pins of the chip and we have full control of the internal facilities of the chip. We work at the grass roots level.

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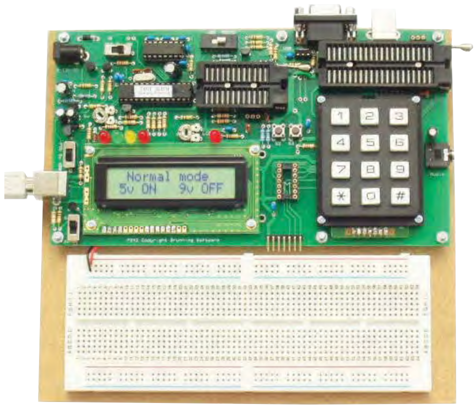

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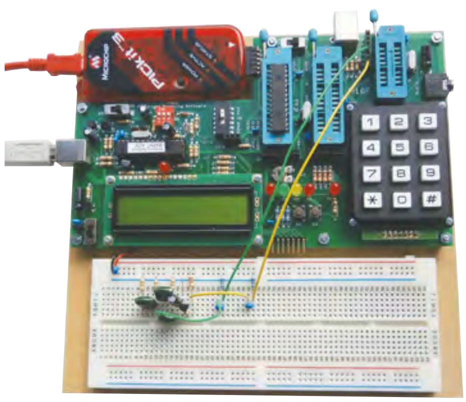
The P942 PIC training course offers exactly the same training as the P931 but the P942 has the ability to verify the data in the PIC at high voltage, normal voltage and low voltage. Verifying that the code in the PIC can be read at high and low voltage is usually recommended if the PIC is to be used in a permanent circuit. For training and experimental work high and low verification is not necessary.



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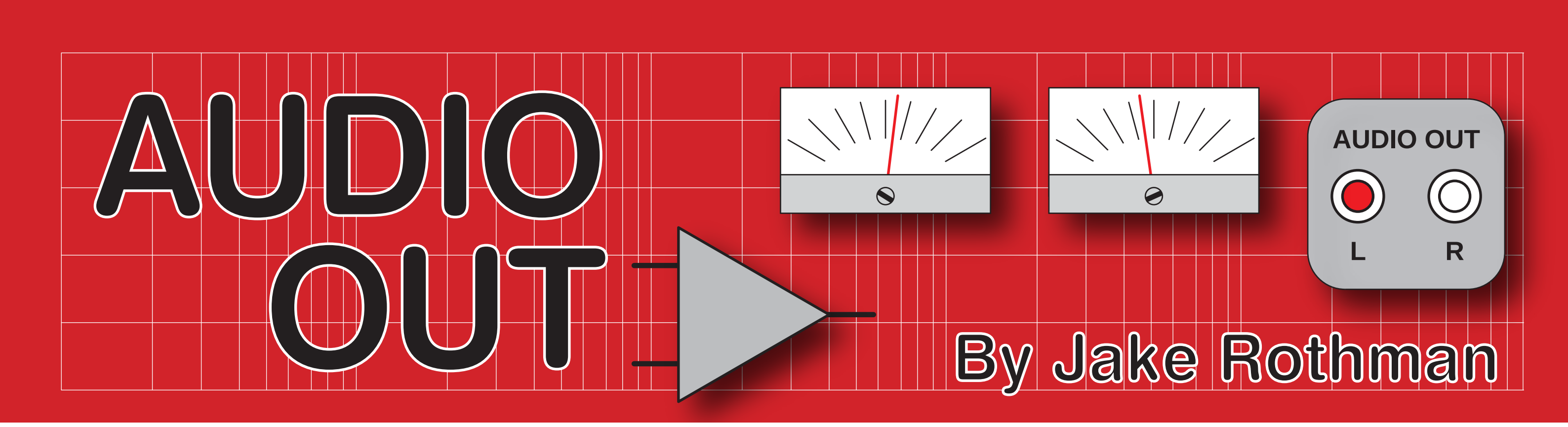


RIAA equalisation – Part 3

Hum and earthing

I often despair with the Hi-Fi fraternity, why do they use unbalanced RCA phono connectors for turntables? In every other area of engineering using a low output floating inductive transducer, balanced lines are used. The result of this heritage issue is that most vinyl playback is accompanied by low-level hum. The bass boost in the RIAA makes it worse by a factor of 10. Move the leads a bit and the hum changes. To further complicate matters there is usually a separate earth lead for the turntable metalwork, as shown in Fig.21. This can be okay, but sometimes it is connected internally to the signal earth, which then gives rise to an earth loop. It is worthwhile trying to connect the earth lead on both the signal earth and chassis earth. A way round the lead problem is to include the RIAA amp inside the turntable, just under the arm. Connecting a turntable would then just be like hooking up a CD player to an amplifier. But that would be too easy and there would be no market for £100 gold-plated phono leads! I have converted decks to balanced operation, but it is essential to separate the cartridge case earth link from the earthy end of the coil. (see Fig.22)

Rumble filtering

A problem that's disappeared with the arrival of digital sources is speaker 'flap' (low-frequency 'wobbling' or disturbances, which reflex speakers are

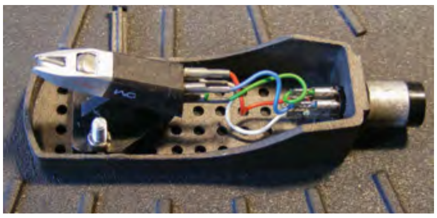


Fig.21. Headshell with Orfoton DJ cartridge showing standard colour coding for the wires. Left live is white, left earth is blue, right live is red and right earth is green. The metalwork for the deck and arm is normally connected via a separate wire to the chassis on the amplifier.

Fig.22. Pin view of Shure M75ED cartridge. Note the earth link from pin to case – this has to be disconnected if balanced operation is to be used.

particularly susceptible to, so reflex speakers are now much more popular). Unwanted low-frequency signals were a real problem with record players, which meant most speakers in the vinyl period had sealed cabinets. These unwanted subsonics arise from the random wobbles in the disc, bearing noise from the deck and thumps through the floor. These disturbances are then amplified through the arm's mass/cartridge compliance resonance, as well as the RIAA bass boost. It is essential to provide some kind of filtering for subsonic signals and to keep the arm resonance in the region of 16Hz, away from record warps around 0.5 to 5Hz, but out of the audio region. This can be checked by tapping the arm while putting it into a digital storage scope (see Fig.23). Electrical high-pass filtering (–3dB at 20Hz) was added as an IEC amendment to the

Fig.23. Oscillogram of Orfoton cartridge-arm combination resonance. A bit high at about 17Hz, but well out of the warp region.

RIAA spec in 1976 and was usually provided by the lower-arm feedback capacitor; 36µF was needed with the standard 220Ω (see R7 in Fig.10, *Audio Out,* August 2015). This was not considered effective enough, so few amplifier designers incorporated it. A better approach is to use a Sallen-and-Key 20Hz third-order Butterworth high-pass filter in the next stage, or an additional network wrapped around the feedback, such as in the discrete pre-amp shown in Fig.24 (C14). The frequency response with the rumble filter is shown in Fig.26. This circuit has a high output drive capability and an alternative low-impedance RIAA network is shown using yet more surplus polystyrene capacitors. Fig.25b. shows the circuit hardwired onto a piece of matrix board.

Cartridge loading

A moving magnet cartridge has a response that falls off due to its inductance and then peaks up at around 18-20kHz where the stylus mass/record interface takes place. This can cause a dip in the upper mid-band response which is audible. By loading the cartridge with a specified capacitance, normally 200 to 500pF with a 47kΩ load, its inductance can be made to resonate at around 9 to 11kHz to give a smooth response up to around 20kHz. Typically, lead capacitance is in the order of 150pF, so 100 to 300pF has to be added to the pre-amplifier input. Ideally, a frequency sweep or pink noise test disc is used to optimise the response, but I've never seen one, all I've found are tracking test discs. I generally put switchable capacitance on the inputs of my pre-amps and tweak to avoid the 6-12kHz dip. Stanton Magnetics have some nice curves showing different loading conditions. (See *The Handbook for Sound Engineers, 2nd Ed*, published by Howard Sams, p.980-2)

Middle and side

One of the interesting things about vinyl subsonic vibrations it that the disturbances result in vertical motion

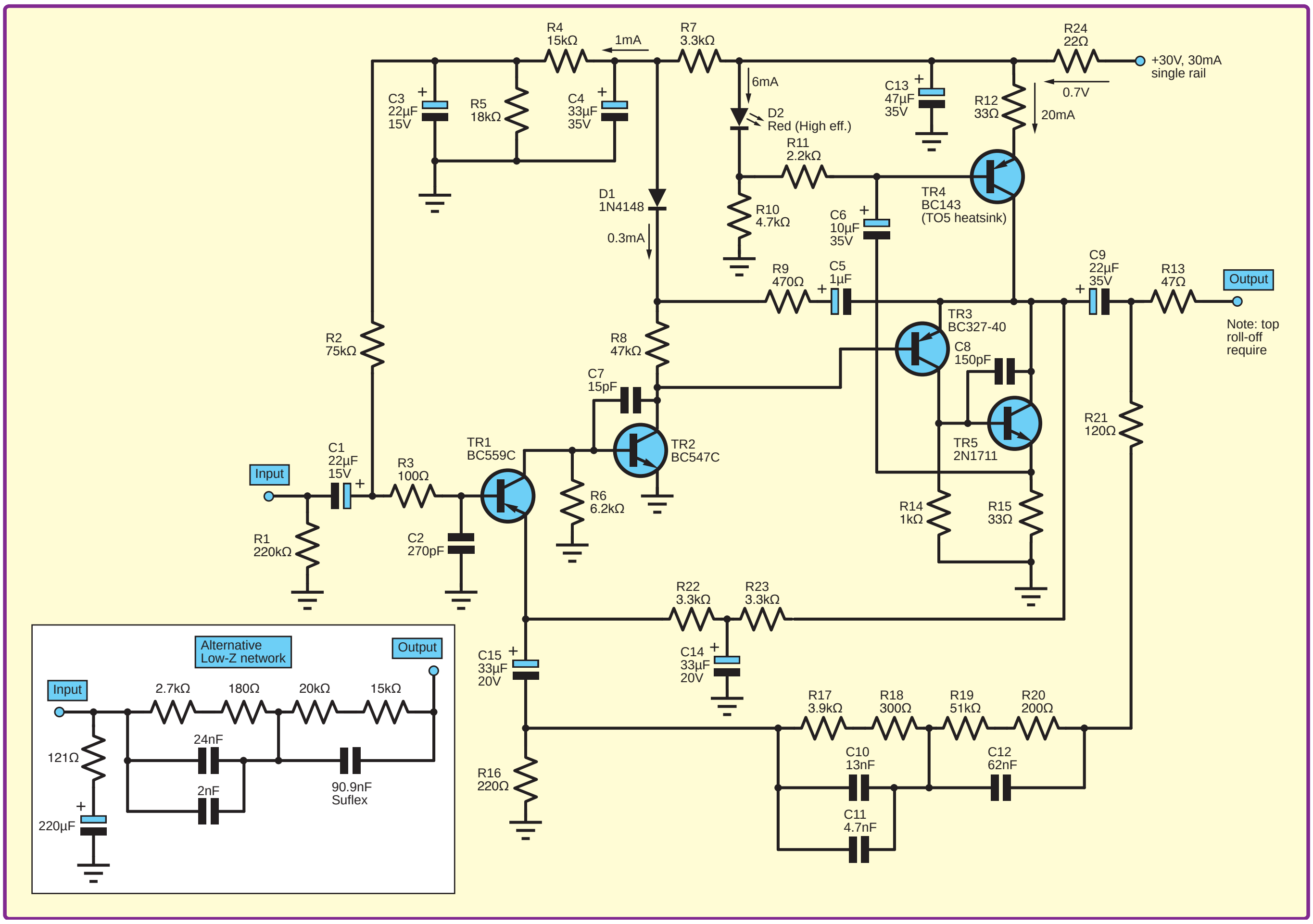


Fig.24. Discrete RIAA amplifier circuit inspired by the 1979 Douglas Self circuit. It can be uprated to higher rail voltages. This design offers better paper specs than op amp circuits, but in practice the benefits are of little audible benefit on real scratchy records (to my ears). This circuit relies on 1.5dB treble cut at 20kHz in later circuits to give a flat RIAA curve.

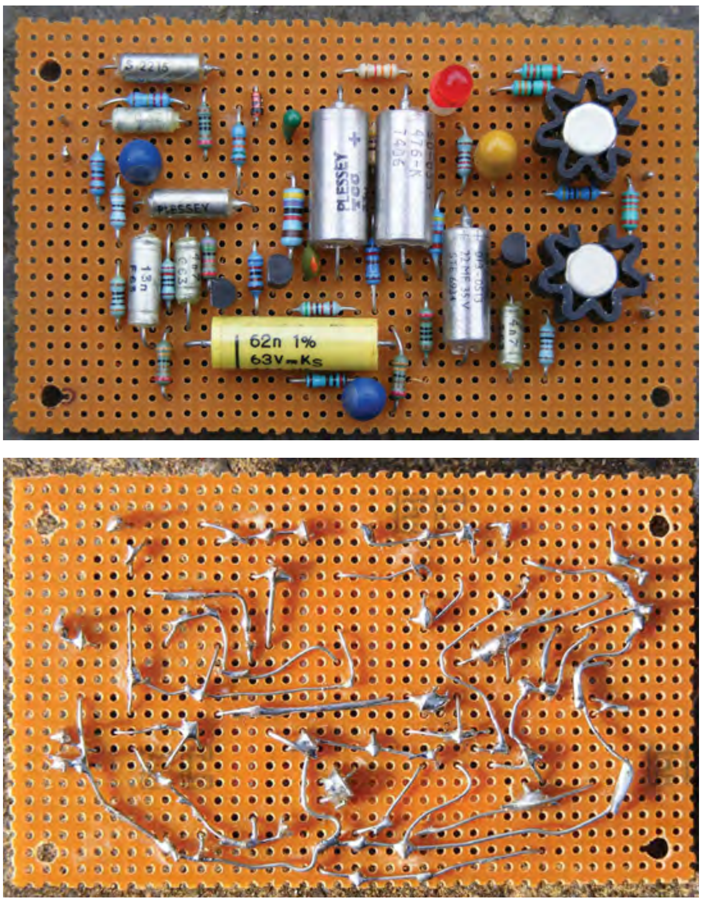


Fig.25 a) (top) Discrete RIAA amplifier built on matrix board, an effective way of planning a later PCB layout (Fig.25 b) (bottom).

of the stylus, which is translated into out-of-phase stereo signals. If you tap a record deck with the stylus resting on a stopped record you will see one speaker cone go out as the other goes in! I used this phenomenon to create a most-effective rumble filter by using the old Blumlein technique of converting the stereo to middle and side (M+S) and high-pass filtering the side/difference signal and then recombining into stereo again. The side signal can also be boosted to increase the stereo separation. The block diagram is given in Fig.27. I plan to provide a full pre-amplifier based on this principle in a later article.

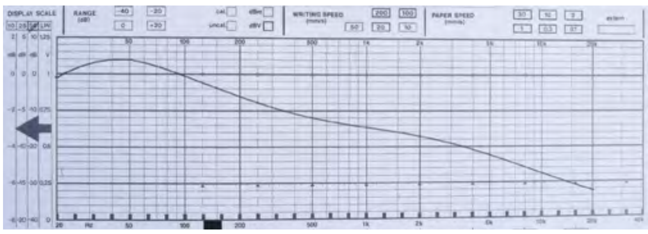


Fig.26. Frequency response with the rumble filter network for Fig.24. It provides some bass boost at 50Hz for my DJ friends!

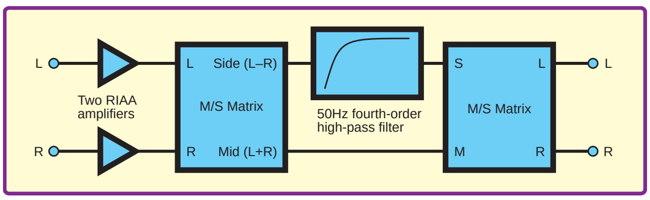


Fig.27. Using middle and side matrixing to get effective rumble filtering with no effect on the bass response.

Max's Hot Beans

By Max The Magnificent

littleBits review

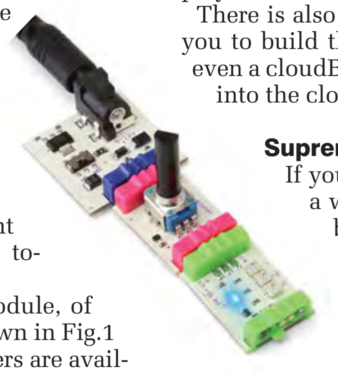
For my birthday one year when I was a kid in the early 1970s, my parents gave me an electronics hobby kit to play with. This involved a lot of discrete components and a lot of springy thingies that were used to connect everything together. (They still make these today, like the Elenco 130-in-1 Electronic Playground and Learning Center, see: **http://amzn.to/1d2pJ5o**)

Unfortunately, although my dad was a wonderful man, he didn't have a technical bone in his body. I did have an uncle, who knew electronics, but (a) he lived 200 miles away in London and (b) he was insane, so I was pretty much left to my own devices. Maybe if I'd had someone to guide me it would have been more fun, but I found it to be horribly confusing. Part of this was due to the manual, which – amazing as it seems – did not appear to have been written with a younger (or even an English) audience in mind. I recall its creators plunging into the topic of 'resistor bridges' without any explanation as to what they were waffling on about. Suffice to say, I was not wearing my happy face.

Now jump forward approximately 45 years (oh my goodness, has it really been that long?). How things have changed. For a start, there are so many cool electronic products around. On the one hand, a lot of things have become both easier to use while being horribly complex 'under the hood' (take GPS, for example); on the other hand, the sophistication of modern components and sensors allows us to create some amazing educational tools.

It's not the size of your bits that counts

Take 'littleBits', for example (**http://bit.ly/18jMIpp**). Created by Ayah Bdeir, these little modules provide one of the best ways I've seen to interest younger kids in electronics and robotics, and to get them up and running as quickly and easily as possible. It's obvious that every aspect of these modules has been painstakingly designed to make them easy to use, from the colour-coded connectors that identify the type of module (blue = power, pink = input, green = output) to the handwriting-style font used for all their annotations.



One great thing about littleBits is that they connect magnetically, and you simply cannot connect them the wrong way round, because if you try to do so they push themselves apart. When you present them in the correct orientation, they snap together with a satisfying 'thunk.'

Every system starts with a blue power module, of which there are various flavors. The one shown in Fig.1 is powered from a wall-wart supply, but others are available, such as one powered by a 9V battery. The simplest system could involve connecting a green output module, like an LED, directly to the power module, in which case the LED will immediately turn on.

Next, we might insert a pink input module between the power supply and the LED, such as the dimmer module shown in Fig.1. Another type of input module might be a push-button switch – press the switch and the LED lights up; release the switch and the LED goes out. This is something even very young kids can wrap their brains around.

There are also logic modules, like NOT, AND, OR, XOR, and so forth. If we were to insert a NOT (inverter) module between the switch and the LED modules, for example, then pushing the switch would now cause the LED to turn off, while releasing the switch would turn it back on again.

The sky is the limit

Although you can start off simply, the littleBits system can be extended to do some amazing things. I just read the book *Make: Getting Started with littleBits: Prototyping and Inventing with Modular Electronics* (**http://amzn.to/1QdJns2**), which was written by littleBits creator Ayah Bdeir (TED Senior Fellow, co-founder of the Open Hardware Summit, MIT 35 under 35) and MAKE editor Matt Richardson. Although this is a small, thin, non-threatening book, it really opened my eyes to some of the incredible possibilities afforded by the littleBits ecosystem. Let's start with the fact that there are more types of modules than you can swing a stick at, and you can even create your own if you so desire.

Of particular interest to me, since I'm currently playing with the Arduino, is the fact that there's even an Arduino Bit. The book provides URLs to a wide variety of user projects. It also provides introductions for a wide variety of kits, all of which you can find on the littleBits website (**http://bit.ly/18jMIpp**).

The clever part is that littleBits has teamed with different companies to form some very interesting things. For example, they partnered with Korg to create the littleBits Synthesizer kit (**http://bit.ly/1cyT3sJ**). This includes modules like an oscillator, random noise generator, keyboard, sequencer, mixer, envelope generator, filter, MP3 player, and so forth.

There is also a variety of wireless modules that allows you to build things like radio-controlled robots; there's even a cloudBit that allows you to connect your projects into the cloud (Internet).

Supreme Commander Max

If you visit the littleBits website, you will find a wide variety of kits, ranging from absolute beginner to 'hunky hero.' The end result is that littleBits is of interest to people of all ages and abilities, from young kids, to high school pupils, to university students, to older folks who have retired and would like to start playing with electronics. I tell you, if I had been given access to the littleBits ecosystem when I was a young lad, I would be Supreme Commander of the World by now!

(Above, left) Fig.1. A littleBits blue power module, pink input module, and green output module

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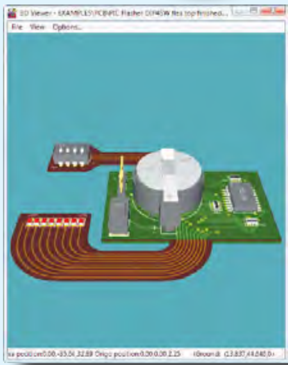

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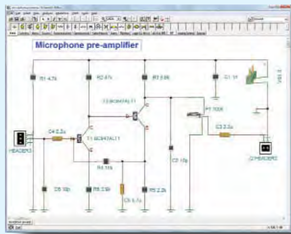

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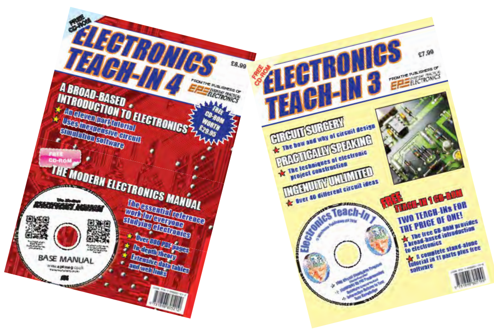

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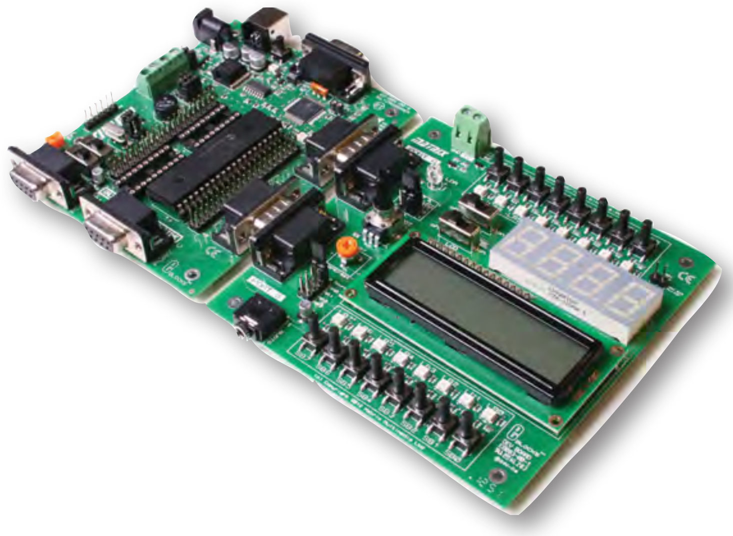

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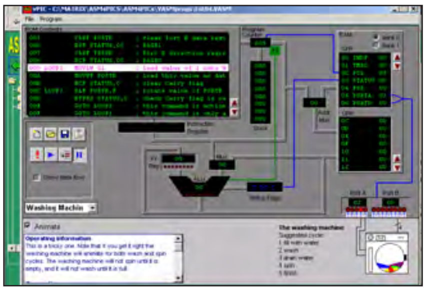

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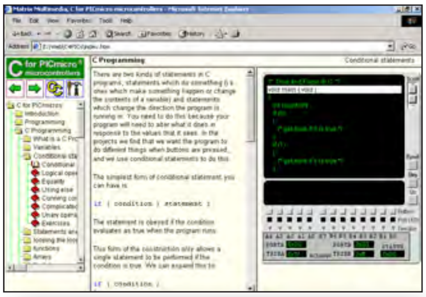
● Complete course in C as well as C programming for PICmicro microcontrollers ● Highly interactive course ● Virtual C PICmicro improves understanding ● Includes a C compiler for a wide range of PICmicro devices ● Includes full Integrated Development Environment ● Includes MPLAB software ● Compatible with most PICmicro programmers ● Includes a compiler for all the PICmicro devices.



Minimum system requirements for these items: Pentium PC running, 2000, ME, XP; CD-ROM drive; 64MB RAM; 10MB hard disk space.
Flowcode will run on XP or later operating systems

FLOWCODE FOR PICmicro V6

Flowcode is a very high level language programming system based on flowcharts. Flowcode allows you to design and simulate complex systems in a matter of minutes. A powerful language that uses macros to facilitate the control of devices like 7-segment displays, motor controllers and LCDs. The use of macros allows you to control these devices without getting bogged down in understanding the programming. When used in conjunction with the development board this provides a seamless solution that allows you to program chips in minutes.

- Requires no programming experience
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- Facilitates learning via a full suite of demonstration tutorials
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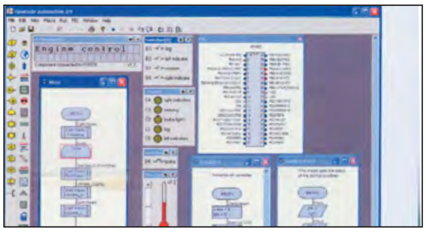
- Pulse width modulation
- I2C.

Please note: Due to popular demand, Flowcode PICmicro, AVR, DSPIC, PIC24 & ARM V6 are now available as a download. Please include your email address and a username (of your choice) on your order. A unique download code will then be emailed to you. If you require the CDROM as a back-up then please add an extra £14 to the price.



PRICES

Prices for each of the CD-ROMs above are:

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(UK and EU customers add VAT to 'plus VAT' prices)

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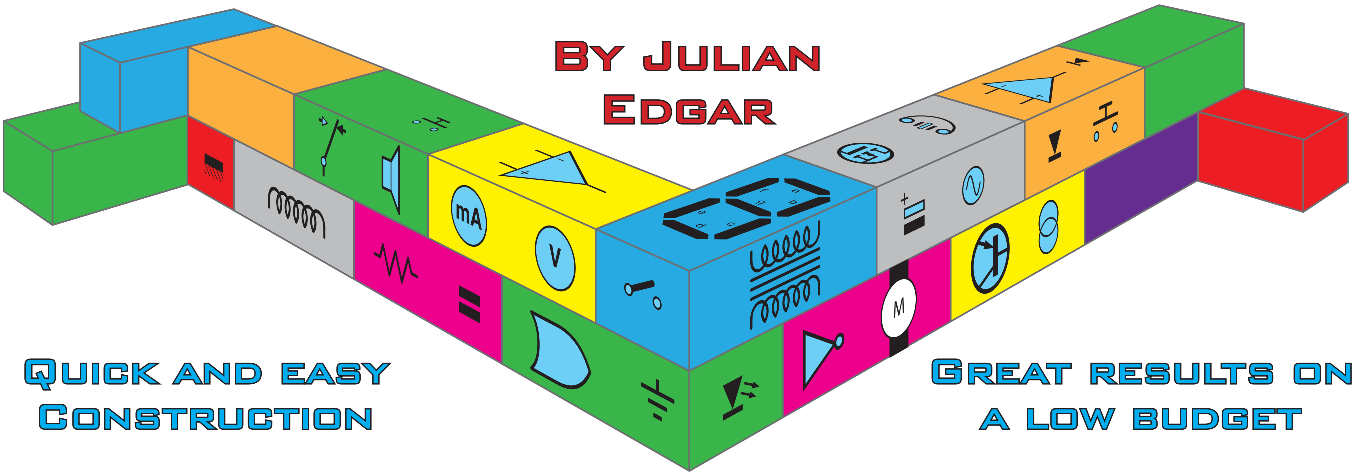
Flowcode (choose PIC, AVR, ARM, dsPIC, PIC24)	**£94.80**	**plus VAT**

Electronic Building Blocks



By Julian Edgar

Quick and easy construction

Great results on a low budget

Temperature Controller

Large complex projects are fun, but they take time and can be expensive. Sometimes you just want a quick result at low cost. That's where this series of *Electronic Building Blocks* fits in. We use 'cheap as chips' components bought online to get you where you want to be... FAST! These projects range from around £15 to under a fiver... bargains!

Temperature Controller and Display

This month we kick off with a brilliant temperature controller that's so cheap it's unbelievable. You want to control a cooling fan based on temperature? Well here is an incredibly cheap 12V module that not only will do that, but also displays the temperature in clear LEDs. How cheap? Try £14 delivered to your UK letterbox. To find it, search on eBay under 'DC 12V Digital Temperature Controller Thermostat C'. At the time of writing a typical unit is eBay number 281081740269 – £13.99 including postage, but hunt around and you my find it for less. Do check that it is the same unit though – or at least does what you want it to do – there are many variants.

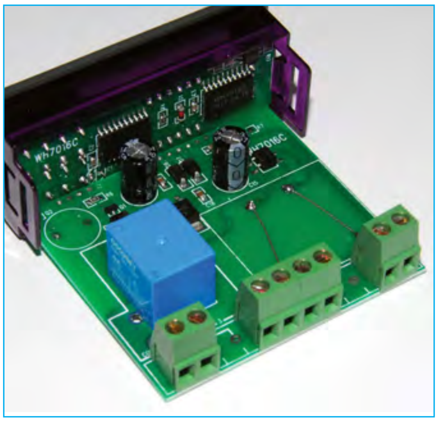
The module is 78 × 71 × 29 mm (L × W × H) and uses a display window that requires a cut-out of 70 × 28mm. It has a mass of 110 grams.

The controller uses an LED display that shows temps up to 100°C to one decimal place (eg, 35.6), and above 100°C in single units (eg, 105). The update rate is fast (about three times a second) and the sensor is very responsive to changes in temperature.

In addition to the numerical display, there are two individual LEDs. One shows when the set-point has been exceeded. (The set-point is the temp at which you've set the device to activate its output.) This LED has two modes – steadily on when the relay is activated, and flashing when the set-point has been passed but the module is running an inbuilt delay before turning on the output. (You can vary this delay time – more on this in a moment.)

The other single LED shows that the display is indicating the set-point temperature.

On the face of the instrument are four push buttons – up/down arrows, Set and Reset.



Inside view, with connectors at the bottom of the photo

LEDs (left) indicate set point and operation

The whole controller and display unit needs just four simple buttons

Making Connections

Wiring connections are by means of screw terminals on the rear of the module.

The module doesn't look at all cheap and nasty – in appearance it could easily be an expensive instrument... and that sentiment also applies to the internal build quality.

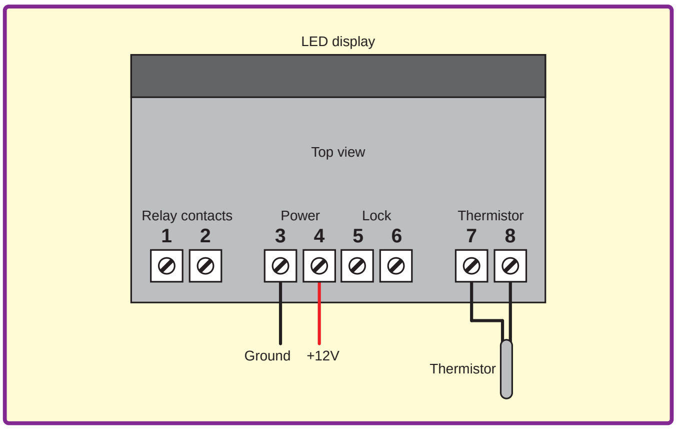


Fig.1. All you need to display just temperature – just four connections

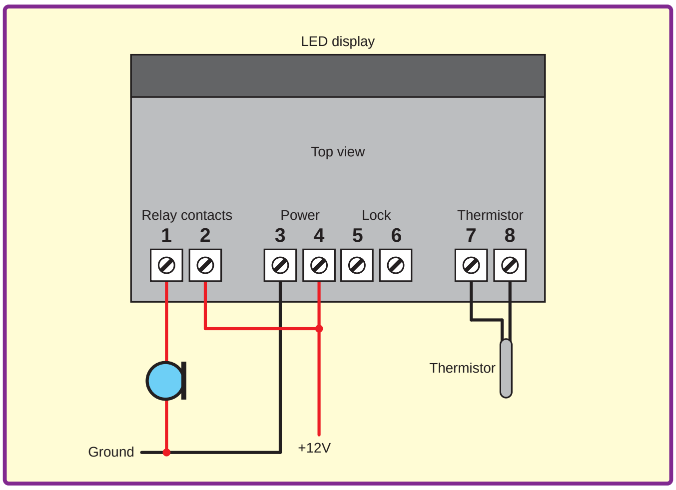


Fig.2. Inside view, with connectors at the bottom of the photo

Displaying Temperature

The simplest use of the instrument is to display just temperature. This requires only four wiring connections and no menu configuration – see Fig.1.

Power – (12V nominal) connects to Pins 3 and 4 – ground to pin 3 and positive to pin 4.

Sensor – the NTC (negative temperature coefficient) sensor that is provided connects to pins 7 and 8 – it doesn't matter which wire goes to which terminal. With these contnections made, the display should come alive and show the temperature at the sensor. Note: the thermistor wiring can be extended as required.

The default values programmed into the instrument mean that straight out of the box it will work fine as a digital thermometer.

Controlling an output

The module is fitted with a 5A relay. This means it can be connected directly to low voltage buzzers, fans and warning lights.

To get a feel for how the control system works, it's a good idea to play with it before installation. Let's take a look at how it can be set up.

Pressing the Set button briefly changes the display to show the set-point temperature. This setting can be altered by pressing the up and down keys. When done, press the Set button again or simply wait a few seconds and the display reverts to the current temperature.

Pressing the Set button for 3 seconds brings up a second menu. Different parameters can be selected by pressing the up/down buttons. To change the selected parameter, press the Set button a second time then make the adjustments with the up/down buttons. Whatever setting is selected is retained in memory, even if power is lost.

The available parameters are as follows:

HC – this menu configures the module to either turn on its relay when the temperature exceeds the set-point ('C' mode), or turns on the relay when the temp falls below the set-point ('H') mode.

d – this sets the difference in temperature between switch-on and switch off. (This is sometimes called the hysteresis.) By using the up/down buttons, you can set this anywhere from 1 to 15°C. This is a very powerful control that can make a huge difference to how the system functions.

L5 – this is the minimum temperature the set-point can be configured. Normally, this would not need to be altered from its –50°C default.

H5 – this is the maximum temperature the set-point can be configured. Normally, this would not need to be altered from its 110°C default.

CA – this function allows you to correct the temperature display by adding or subtracting 1°C from the displayed reading.

View of the units connections

P7 – this function is used when in 'C' mode you don't want the output cycling on and off at short intervals. The setting can be anything from 0 – 10 minutes. For example, if it is set to one minute, after the relay has activated once, it will not activate again until a minute has passed – even if the temp set-point has been tripped. In most uses you would set this to zero.

The wiring for an over-temperature alarm buzzer is shown in Fig.2.

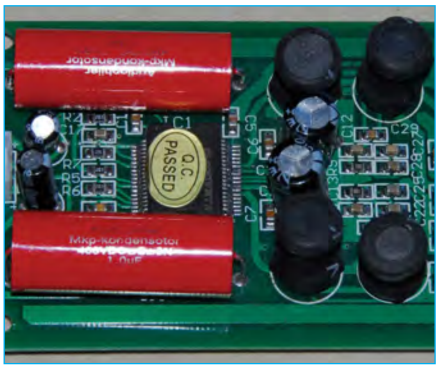
Conclusion

For the money, this is just a phenomenal instrument. Having an accurate and fast-response digital thermometer is just great – but being able to smartly control a relay output is just amazing. The icing on the cake is the adjustable hysteresis and delay to prevent cycling. Oh yes, and the fact you can calibrate it… I think you will agree, what a superb piece of kit for the price!

Next month

Need a cheap audio amplifier – perhaps as a test bench amplifier or for a kid's MP3 sound system? Well it doesn't get much cheaper than this – try under £4 for the prebuilt amplifier module, including postage to your UK letterbox!

Tune in to our next *Electronic Building Block* article.



Next month's amplifier

READOUT

Matt Pulzer addresses some of the general points readers have raised. Have you anything interesting to say? Drop us a line!

All letters quoted here have previously been replied to directly

Email: editorial@wimborne.co.uk

☆ LETTER OF THE MONTH ☆

50 years a reader!

Dear editor
Your recent retrospective of the 2N3055 power transistor calls to mind another (and very early!) episode in my 'career' as an electronics enthusiast. My senior schooling was at Northgate Grammar in Ipswich, and a previous student project had been the construction of an electronic gymnasium scoreboard for basketball or the like. I was asked if I would like to finish it off.

Readers need to understand that this was the late 1970s: electronics was by no means as ubiquitous as it is now, the range of devices obtainable and affordable was limited*, and the LEDs of the day were dim and expensive. Consequently, to construct a large-format four-digit display (0-99 for each of two teams), incandescent pygmy lamps were used to create seven-segment digits – from recollection, 23 lamps per digit.

The project I inherited used a bank of 28 2N3055s (one per segment) to drive the lamps, interfacing four BCD thumb wheel switches (similar to those used in the capacitor section of the recent decade box project) via BCD to 7-segment decoder chips (TTL). Parts were, I believe, donated by the Post Office research labs at Martlesham Heath.

Unfortunately, the power drivers were (of course) limited to DC, and 60V, so the main supply rail was 50V from a hefty transformer via a bridge rectifier, and the lamps were 50V types. Even at 15W per lamp, the total possible load was over 400W and 8A – the transformer was simply not big enough.

In order to try to rescue the situation, I converted to thyristors controlling mains. Triacs were still unheard of, or at least too expensive, and the early ones needed bipolar triggering anyway. This gave me 240V half-wave, equivalent to about 120V, and I wired the lamps in each segment in series rather than parallel. This worked, but the lamps were now under-driven, showed a degree of flicker, and because some segments had four lamps and others three there was a brightness difference between segments.

The next thing I tried was rewiring the lamps to parallel again, and using phase control on the thyristors to reduce the mains power. This was very successful... until a lamp blew. The brief current spike that an incandescent lamp produces when it expires is too brief to blow a fuse, but it's plenty long enough to blow a thyristor!

That, as I remember it, is the way the project remains. I wrote it up and it was published in *ETI*. The unit is about 4-feet × 18-inches × 12-inches and remains in my mother's loft in Ipswich. To get it working reliably, I would probably swap to triacs with unipolar triggering, and substitute 240V lamps for the 50V ones (too expensive at the time). It might even be practical to use 12V LED lamp substitutes intended for automotive use and go back to DC.

I'm sure my mum would be delighted to have this out of her loft space if anyone would like to take it away. Do we have any readers from Northgate, or other not-for-profit organisation in that area?

* It seems to me the situation is not much better now. Through the hay-day of the 1980s, the amateur enthusiast could tap into the mass producers supplying professional industry; there was a huge range of standard logic devices available in normal through-hole packages suitable for the home hobbyist to assemble on Veroboard or home-made PCBs. The professional world has moved on, with microscopic surface-mount parts and custom silicon, so the availability of parts suitable for hobby use has shrunk.

Ken Wood, by email

Matt Pulzer replies:
Thank you for your reminiscences Ken, it sounds a little like a project I built at school, albeit much more sophisticated than my attempts. I do remember asking my physics teacher how to design with transistors and being somewhat stumped with load lines and the whole non-linearity of his explanation. In retrospect, I now realise that what I really wanted – without realising it – was a nice little package like an op amp that I could string together with other devices like electronic Lego. Eventually, my teenage brain got the hang of transistors, but it did take a bit of head scratching.

Cheap as chips

Dear editor
With reference to the *Ge-Mania* article in the June 2015 issue of *EPE*, Jake Rothman sets an interesting question as to whether the Sinclair 750 circuit is correct.

First, let me say that many of Sir Clive's circuits were built for the lowest cost possible. Many in the industry criticised them for being pared too far: a few pence more on one component or other would have made the products industrially worth considering. So it is with the 750 – it could hardly have been any cheaper.

A quick evaluation of the circuit and its claimed performance suggests that as drawn there is something wrong. However, there are some important considerations. First, using a loudspeaker as a load means that a resistive load may not be accurate.

For example, a 30Ω load is unable to develop 750mW RMS from a 12V supply, but if this were peak, 375mW is more likely. Even this is dubious, but without a suitable loudspeaker model it is impossible to say. At best, a 1:1 isolation transformer could yield 0.9W RMS, but only if the quiescent current were 300mA, requiring 3V across the emitter resistor, peaking at 6V. So it would appear that 750mW is not the RMS at all, but even 375mW seems optimistic.

In principle, the bias current of the output transistor would be in the region of 20mA if the input transistor were

an ordinary germanium type, such as the AC126 or OC71. However, the MAT120 was a high frequency device and may have had a low gain. If Sinclair used rejects, it could have possibly been vey low, such that a high voltage drop was needed on the base bias resistor, thus giving a high voltage across the 10Ω emitter resistor, conceivably putting the output quiescent current up to 100mA or more. With any load, though, 100mA is only 1.2W input, so even then we do not get 750mW unclipped output, only 600mW!

The bandwidth stated seems unlikely with the circuit as drawn. But here also, the transistor characteristics are critical. The MAT120 may have had a high input impedance that would have allowed the 100nF input capacitor to provide a 30Hz LF cut-off; but with standard transistors like the AC128 – or even a BC558 – the input capacitance would have to have been much higher, say 2.2µF.

To summarise, the answer to Jake's question – if the circuit is correct as drawn – is 'probably', but this depends critically on the actual characteristics of the transistors and loudspeaker specified. Given the intended use of the circuit, which was to provide a loudspeaker output from a Micro 6 receiver so that you did not need to use earphones, it probably just meets its expectations, and if so I would conclude that the transistor characteristics may have been quite odd compared with standard devices.

What we can say is that with more modern transistors, the circuit would not meet the stated specifications, and would need some significant modifications. Certainly it would seem that a distortion figure would be around the 2% mark, and my interpretation of the statement 'it's Hi-Fi in miniature' would be that it has very little Hi-Fi associated with it!

However, neither the output power conditions, nor the distortion figures are forthcoming, making their statements difficult to conclusively challenge.

It is worth stating, perhaps, that the classic two-transformer circuit would have been a much better bet. These amplifiers tended to need only 2 or 3mA in the driver stage and maybe 2mA queiscent current in the output. Thus, the quiescent current was lower than in many complementary circuits!

John Ellis, by email

Jake Rothman replies:
I thank John for his insight and theoretical analysis of Sinclair's TR750. It's a shame we can't get an actual one to test – does anyone out there have an example I could examine?!

I don't think there's too much of a problem defining the speaker load, since every drive unit I've measured has a DC resistance of around 80% of its stated impedance. In the case of 30Ω, we can be pretty sure the DC resistance will be around 24Ω. In conjunction with the emitter resistor, peak current would be limited to around 350mA.

The input capacitor was probably made small in order to save money and provide a 'louder' high-pass filtered sound for radio.

I suspect the TR750 never really worked properly, like most of the Sinclair stuff I've had! I remember a Micromatic radio kit I had as a kid in the early 1970s. It used a trimmer capacitor as the main tuning capacitor. I later found the data sheet that said the device had a maximum life of 10 rotations!

Geo-phys request

Dear editor
I have been a subscriber to *EPE* for some time, but there is one topic that I think you have not covered. Please can you provide a full circuit/design for building a geophysical inspection system for archaeology?

I am a committee member of the UCCA (Upper Coquetdale Community Archaeology) group, which comprises quite a few retired archaeologists and some who are still working. I had not worked with archaeologists previously, but I now realise that it is a very interesting occupation.

Peter Robinson, by email

Matt Pulzer replies:
Dear Peter – that's a great idea, although it may take some time to find somone who can put one together for us. There do seem to be quite a few systems out there, from electrical resistance meters and metal detectors through to ground-penetrating radar, magnetometers and electromagnetic conductivity systems. Some of these might be feasible… others may not fit within the constraints of what is possible on a hobbyists budget. No promises, but I will see what I can do. In the meantime, is there a reader who can help us create some 'smart diggers'?

Ingenuity Unlimited?

Dear editor
While having a spring-clean I noticed that you seem to have stopped running *Ingenuity Unlimited*. I enjoyed *IU* articles and thought they were a great way to stimulate creativity and receive new ideas from your readers. I'm not sure why the series was dropped, but could it be reinstated or a similar reader creativity/innovation column introduced?

I think it would be splendid if *EPE* could, once again, help stimulate creativity in your young readers and invite contributions.

Martin Roantree, by email

Matt Pulzer replies:
Martin, you are right, IU *was/is an excellent series. It hasn't actually been 'dropped', but (temporarily) crowded out by other material. I do hope to find space for its revival in the near future – thanks for the reminder!*

Analogue computing

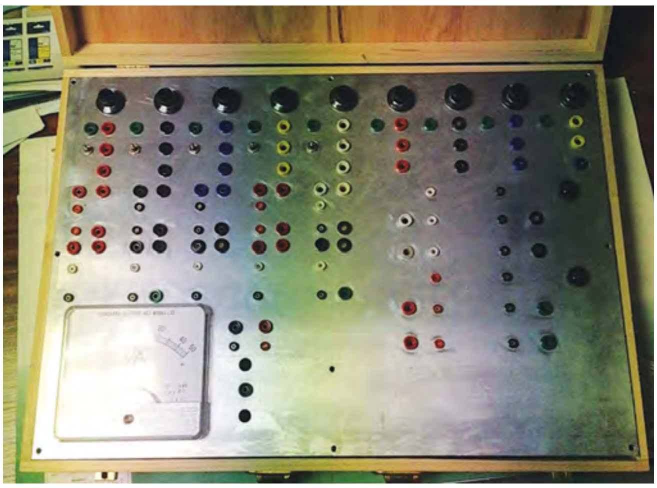
Dear editor
I read with interest Tuck Choy's email in the April 2015 issue. I built a PEAC analogue computer in 1968 and still use it to this day. Over the years it has been improved with regards to power supply, operational amplifiers and integrator, plus all peripheral units are now on one panel, which fits into a case. However, the co-efficient potentiometers and voltage source op amps layout is still to the original front-panel design (see photograph).

Over the years I have built some excellent projects by your authors – to mention just a few from the 1960s: Inexpensive Oscilloscope, P Cairns 1965; Transistor & Diode Tester, B Crank 1968; and Investigator Oscilloscope, R Hirst 1967.

These have given me excellent service but only the PEAC is still with me. I am coming up for retirement and I shall remain a stalwart reader of your excellent magazine. Keep the projects coming!

Brian Francis, by email

Matt Pulzer replies:
Great to hear from you Brian and I hope retirement brings you the time to indulge yourself in more electronics fun. We would love to hear how you use your analogue computer and perhaps see some photos of its innards!



PCB SERVICE

CHECK US OUT ON THE WEB

Basic printed circuit boards for most recent *EPE* constructional projects are available from the *PCB Service*, see list. These are fabricated in glass fibre, and are drilled and roller tinned, but all holes are a standard size. They are not silk-screened, nor do they have solder resist. Double-sided boards are **NOT plated through hole** and will require 'vias' and some components soldering to both sides. * **NOTE: PCBs from the July 2013 issue with eight digit codes** have silk screen overlays and, where applicable, are double-sided, plated through-hole, with solder masks, they are similar to the photos in the relevent project articles.

All prices include VAT and postage and packing. Add £2 per board for airmail outside of Europe. Remittances should be sent to **The PCB Service, Everyday Practical Electronics, Wimborne Publishing Ltd., 113 Lynwood Drive, Merley, Wimborne, Dorset BH21 1UU. Tel: 01202 880299; Fax 01202 843233; Email: orders@epemag.wimborne.co.uk. On-line Shop: www.epemag.com**. Cheques should be crossed and made payable to *Everyday Practical Electronics* **(Payment in £ sterling only)**.

NOTE: While 95% of our boards are held in stock and are dispatched within seven days of receipt of order, please allow a maximum of 28 days for delivery – overseas readers allow extra if ordered by surface mail.

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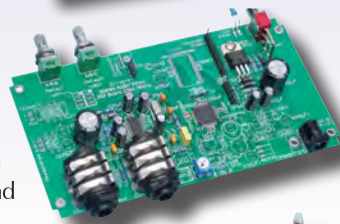
Next Month

Content may be subject to change

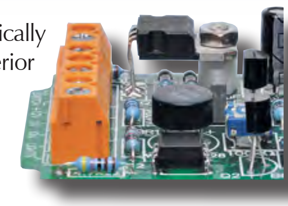
Passive Direct Injection Box

Hum and noise plaguing your performance? Then add this vital piece of equipment to your musician's or roadie's toolbox. Our *Passive Direct Injection Box* converts unbalanced signals from a musical instrument into a balanced output. It performs as well as a powered unit in many applications and doesn't require batteries.



Digital effects processor for guitars & musical instruments

This deceptively simple unit provides 10 different musical instrument effects, including echo, reverb, tremolo, fuzz, compression, flanging and phasing. Each effect is adjustable and can be defeated with a foot pedal switch. It's designed for use with electric guitars, but will work with other instruments and vocals too.

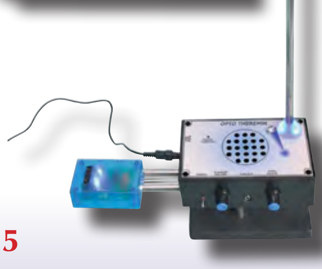


Courtesy LED Lights Delay For Cars

Most modern cars have a courtesy light delay, but older vehicles do not. This new circuit is specifically designed to suit LED lamps, but will also work with conventional filament lamps. It keeps the interior lights of your car lit for a preset time after you shut the car doors. The lights will also turn off if the exterior lights or ignition are switched on during the time-out period.

Opto-Theremin – Part 2

In Part 1 we described how the *Opto-Theremin* works and gave the assembly details for the two PCBs. Next month, we complete the construction, and describe the test and adjustment procedure.



Teach-In 2015 – Part 9

October's *Teach-In 2015* will examine the practical aspects of measurement, adjustment and fault-finding in power amplifiers. Plus, we will look at stability, along with thermal and over-current protection. Our final practical project will be devoted to a low-cost high-quality 10W power amplifier that will out-perform most of today's integrated circuit amplifiers.

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